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MAINS-VOLTAGE STABILIZERS

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Many types of electrical apparatus require a power supply with a very stable voltage. As the mains seldom satisfies this requirement, many types of auxiliary apparatus have been developed during the last two decades, to stabilize the mains voltage to a more or less constant value. This article deals with some of the existing types, and in particular with a new apparatus with some interesting improvements, which has been developed at the Philips X-ray Laboratory at Balham (London).

Introduction

It is well known that the voltage and frequency of the mains are subject to continuous fluctuations. In applications for which these fluctuations (and especially those of the voltage) are too large, matters can be improved by the use of a mains-voltage stabilizer. The requirements to be met by such apparatus may vary considerably according to the particular application. The principal requirements are discussed below.

a) The output voltage of the stabilizer should be as far as possible independent of the input voltage. The "factor of improvement", i.e. the ratio of the voltage fluctuations at the input to those at the output, should be very high for certain purposes, often as much as 100 (e.g. for the calibration of kWh-meters). In other cases a factor of 10 will be ample. An example of this is the stabilizing of the filament voltage of electron tubes, which is worth while for large installations as this results in a longer working life of the tubes.

b) The output voltage should not be dependent on the mains frequency. Although the frequency of most lighting mains averaged over a few hours is kept reasonably constant (in view of the use of synchronous clocks), short-term fluctuations of a few cycles per second may still occur.

c) Some applications require that the variations in the output voltage caused by variations in the load or load power factor, remain small. Where the load remains constant, this requirement is of course unnecessary.

d) Some mains-voltage stabilizers produce an output voltage which is by no means perfectly sinusoidal. For some purposes this cannot be permitted (e.g. in the calibration of A.C. instruments).

In this connection it may be noted that in most cases (e.g. for the heating of filaments) a constant *r.m.s.-value* of the voltage is required. Even if this last requirement is satisfied, a considerable distortion of the waveform of the output voltage (and also a fluctuation of its peak value) can still occur due to the presence of harmonics. If the load consists of a rectifier this is impermissible, since the rectified voltage is often a function of the peak value rather than of the r.m.s. value of the alternating voltage.

e) A further criterion in the choice of a stabilizer is the speed of regulation. If the mains voltage of the load fluctuates in sudden jumps, the output voltage will, as a rule, deviate for a short time from the desired value. The time necessary to reduce this deviation to 1/e of its maximum value, is called the recovery time. In some cases this recovery time is not critical, e.g. in the

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stabilization of the filament voltage and high tension of an X-ray tube used for making long exposures. Here we are concerned only with the total X-ray dose (i.e. integrated over the time). If, however, short exposures are to be made, then the recovery time of the stabilizer has obviously to be short too.

The large number of different types of stabilizer now commercially available is due to the fact that satisfying all requirements at the same time is difficult and therefore often uneconomical; for this reason it is usual to design a stabilizer so that it meets only those requirements essential to a given application.

After a brief description of two types of stabilizer now in common use, a more thorough treatment will be given of another system, developed by the Philips X-ray Laboratory at Balham (London) which combines the advantages of both these types.

"Magnetic" stabilizer

For the "magnetic" stabilizer use is made of the non-linear relationship between magnetic induction (B) and field strength (H) in ferromagnetic materials. For a soft-iron core, for example, this results in a similar non-linear relationship between the applied alternating voltage (V) and the alternating current (I) passing through the choke, as indicated in *fig. 1*. The field strength is, in fact, proportional to the current, whereas the voltage is determined by $\mathrm{d}B/\mathrm{d}t$. It is obvious that, if the voltage is sinusoidal, the current will not be, and vice versa. As regards the r.m.s. value of voltage and current, however, the relationship of fig. 1 applies.

Roughly speaking, the choke can be said to behave as a self-inductance that becomes smaller as a higher voltage is applied to it. Such a choke shunted with a capacitor forms a resonant circuit, which, if fed with an alternating voltage of constant frequency, will resonate at a certain value of the applied voltage. This is illustrated by *fig. 2*. Here

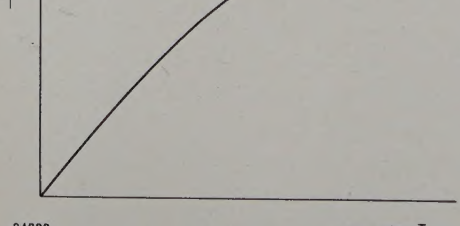

Fig. 1. Relationship between r.m.s. current and voltage for a choke with soft-iron core.

again the relationship between the current through the choke and the applied voltage is given by curve *1*. The straight line (*2*) shows the same relationship for the condenser current. The latter is in anti-phase

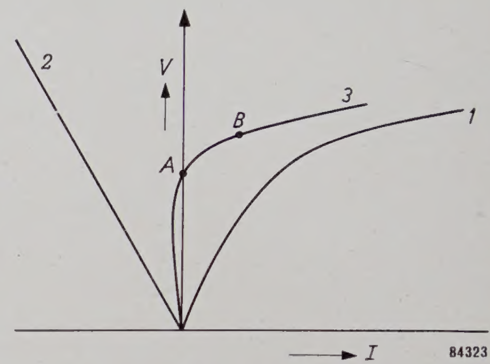

Fig. 2. Relationship between current and voltage for *1*) choke (as in fig. 1); *2*) condenser; *3*) choke and condenser in parallel.

with the choke current, so it has been drawn in the negative x-direction. The sum of both currents (curve *3*) passes through zero at a certain value of the applied voltage (point A). Here the circuit is in resonance. As the voltage becomes higher than the resonance value, the current consumed by the resonant circuit will rapidly increase. Conversely, we may express this by saying that in this range (about point B) the voltage across the resonant circuit is only slightly dependent on the current flowing through it. A resonant circuit of this kind, therefore, forms an excellent stabilizer when connected in series with an unsaturated and hence linear choke, which governs the order of magnitude of the current such that the resonant circuit adjusts itself to a voltage just above resonance (*fig. 3*,

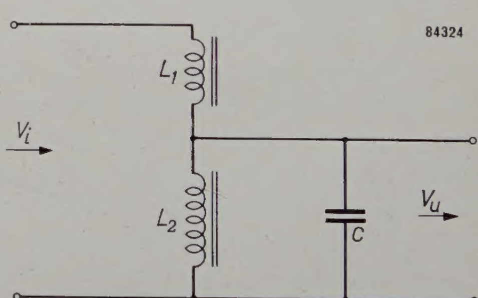

Fig. 3. Simple magnetic stabilizer. V_i = input voltage. V_u = output voltage, L_1 = non-saturated choke, L_2 = saturated choke. The resonant circuit L_2C is in resonance at one particular voltage.

operating point B of fig. 2). In this argument, hysteresis and other losses are neglected.

Without the condenser a stabilizing action would still remain. The difference between curves *1* and *3* of fig. 2, however, shows that the flat portion of the V-I relationship, necessary for good stabilizing, will occur at a smaller current value if a condenser is used.

Because of the fact that the voltage across the resonant circuit is so little dependent on the current passed through it, the output voltage is likewise little dependent on the current extracted from the output terminals.

Several improvements may be incorporated in the circuit of fig. 3. First of all an auto-transformer may be used for L_2, so that the output voltage V_u becomes as high as the nominal value of the input voltage V_i. L_1 may further be provided with an additional winding (*fig. 4*). The voltage V_1 produced by this winding is small compared with V_2 and of opposite polarity. The remaining small variations in V_2, caused by fluctuations in V_i, will then be compensated by the variations in V_1.

In the above description we have not considered the various complications that may occur. For a more thorough analysis reference can be made to the literature on this subject [1]).

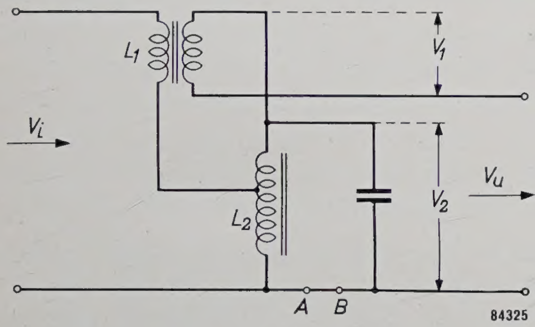

Fig. 4. Stabilizer similar to that of fig. 3, with compensation winding on L_1. Here L_2 is in the form of an autotransformer, so that V_u is equal to the nominal value of V_i.

Another version of this type of stabilizer employs one single iron core (*fig. 5*). The magnetic flux

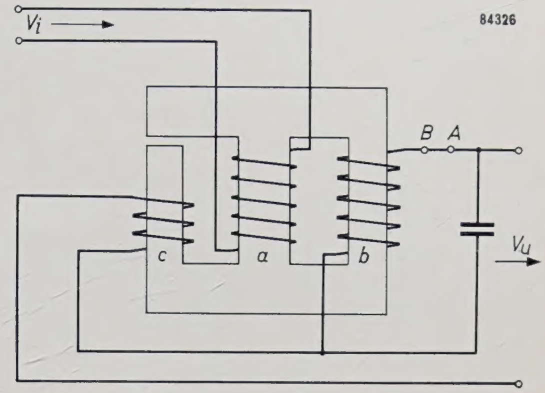

Fig. 5. Magnetic stabilizer with one core. The right limb becomes saturated earlier than the left limb with the air gap.

caused by the winding *a*, is distributed between the two outer limbs of the core. If the voltage V_i across *a* is increased, the greater part of the flux will initially pass through the right limb, because on the left the path is interrupted by an air gap. The right

[1]) W. Taeger, Spannungsgleichhaltungs-Schaltungen mit Eisendrosseln, Arch. Elektrotechn. **36**, 310-321, 1942. R. O. Lambert, Voltage regulating transformers, Electronic Engng. **15**, 384-387, 1943. H. A. W. Klinkhamer, Equivalent networks with highly-saturated iron cores with special reference to their use in the design of stabilizers, Philips tech. Rev. **2**, 276-281, 1937. and **6**, 39-45, 1941.

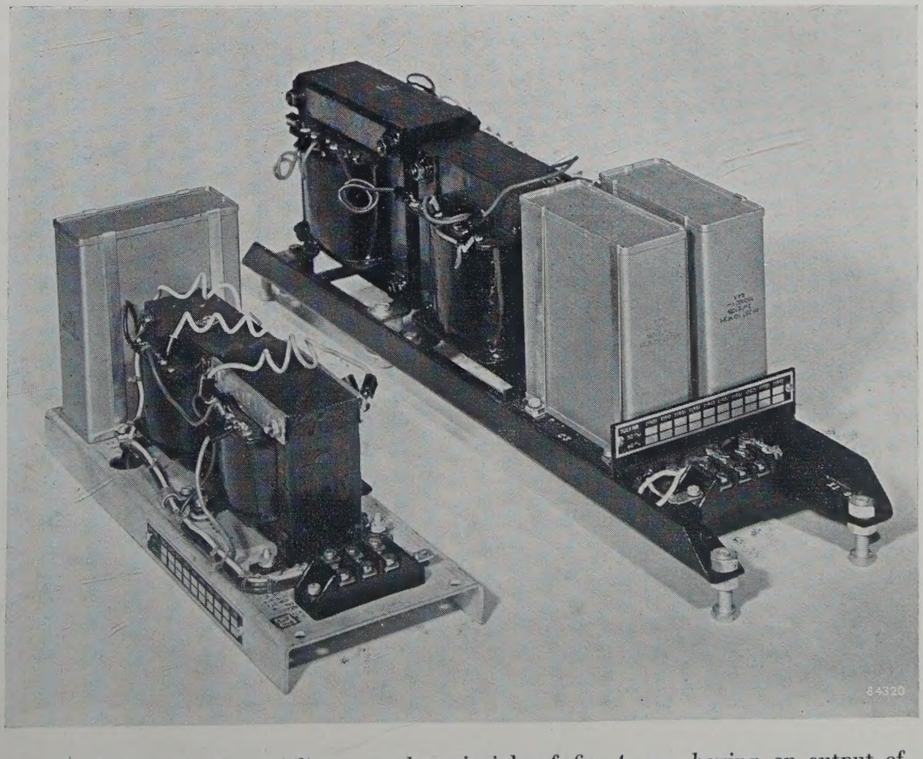

Fig. 6. Two magnetic stabilizers on the principle of fig. 4, one having an output of 80 VA and the other of 300 VA.

Table I

Stabilizer according to	Power output (VA)	Variations of output voltage for				Response time (seconds)
		10% variation of input voltage	5% frequency variation	Load variation between 10% and full load		
				$\cos\varphi = 1$	$\cos\varphi = 0.75$	
Fig. 6	80, 300	< 0.2%	< 8%	< 1%	< 6%	0.02-0.04
Fig. 9	1000, 2000, 5000	—	—	< 0.2% (input voltage and frequency variations included)	< 0.5% (input voltage and frequency variations included)	0.1-0.2
Fig. 16	150	< 0.1%	< 0.2%	< 0.2%	< 0.2%	0.02-0.04

limb, however, is thinner than the central one and thus becomes saturated at the higher values of V_i. Any further increase of V_i then cannot cause much change in the flux through the right limb. Any further increase of the flux through *a* will find its way through the left limb.

The effect of the capacitor in parallel to the *b* winding is analogous to that of fig. 3, although this is not at once apparent. Winding *c* again compensates for the remaining influence of V_i on V_u, and raises the factor of improvement. This winding may also be situated on the central leg.

The circuits of figs. 4 and 5 are both sensitive to frequency changes. In order to keep the oscillatory circuit in the region of resonance, it is necessary that at increasing frequency the self-inductance of the "saturated" coil becomes smaller. This means that the current through the coil and hence also the voltage across it, will increase. With these stabilizers a frequency change of 1% as a rule causes a variation of 1.5 % in the output voltage. Changes of load power factor also have a considerable effect on the output voltage. The regulating speed is high; a sudden voltage surge is compensated within one or two cycles. By choosing the right number of turns for the compensation winding, it is possible to keep either the r.m.s. value, or the mean value, or the peak value of V_u constant. The shape of V_u is not sinusoidal, but depends on V_i and on the load current.

Fig. 6 shows two stabilizers, one for 80 VA and one for 300 VA (with circuits based on the principle of fig. 4). In *Table I* some of their properties are listed.

Stabilizers with feedback

The method to be discussed here (*fig.* 7) uses a detecting device *B* for determining the deviation of V_u from the nominal value. Between the input and output voltage of this detecting device there is a relationship as shown in *fig. 8*. V_1 is zero if V_u has the nominal value V_{u0}. This "error signal", after being amplified in *A*, is applied to a regulator *C* which adjusts V_u.

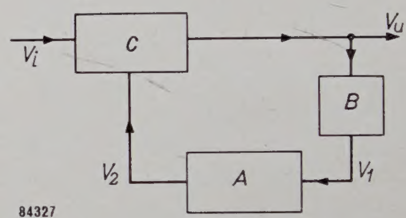

Fig. 7. Block diagram of a stabilizer with feedback. The deviation of V_u from its nominal value is converted into a voltage V_1 by the detecting device *B*. This error voltage is amplified by *A* and is used for regulating the output voltage by means of the controller *C*.

It will be clear that the constancy of V_u improves the more the error signal is amplified. In fact, it can be demonstrated than the factor of improvement (see *a*, p. 1) is equal to the overall amplification +1, the overall amplification being the total amplification in the closed circuit.

The non-linear element used to obtain a characteristic such as that shown in fig. 8, may take various forms. As a rule it has to react to the *r.m.s.-value* of V_u. In view of the definition of r.m.s.-value, this means that the output voltage of the controlling device should be a measure of V_u^2,

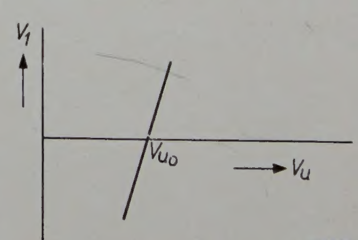

Fig. 8. Relationship between V_u and V_1 (fig. 7). V_1 is a measure of the deviation of V_u from its nominal value V_{u0}.

averaged over a certain time (at least one period). As a rule use is made of the quadratic relationship between the current through a resistor (proportional to V_u) and the power dissipated in it. The resistor temperature is a measure of the r.m.s. current; the averaging function is performed by its heat capacity. To ensure that the instantaneous current values over at least a whole cycle contribute about equally to the result of the averaging, the heat capacity should be so great, and/or the heat exchange with the surroundings so small, that the current (and therefore V_u^2) is in fact effectively averaged over a number of cycles. The delay produced in this way is characteristic of all mains-voltage stabilizers with feedback.

The stabilizers (5, 2 and 1 kVA) shown in *fig. 9* have a detecting device whose non-linear element consists of a diode operating in the saturation range. The filament of this diode (B_1, *fig. 10*) is connected, via transformer T_2, to V_u. The anode current is now highly dependent on the temperature of the cathode, and because the diode forms one of the branches of a D.C.-fed Wheatstone bridge, there exists a relationship similar to that of fig. 8 between V_u and the output voltage of the bridge. In this arrangement it is the filament that forms the squaring, averaging and delaying element.

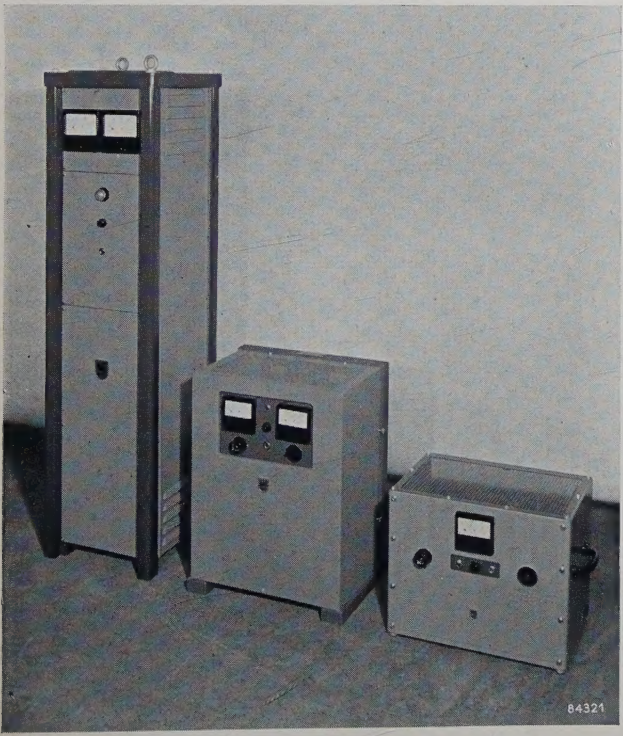

Fig. 9. 5, 2 and 1 kVA stabilizers on the principle of fig. 7.

The regulating part consists of the transductor *Td*, which is a choke whose self-inductance can be varied by passing a direct current through one of the windings. The step-up ratio of T_1 is in this arrangement a function of the self-inductance of *Td*. Here too, therefore, the saturation properties of soft iron are employed.

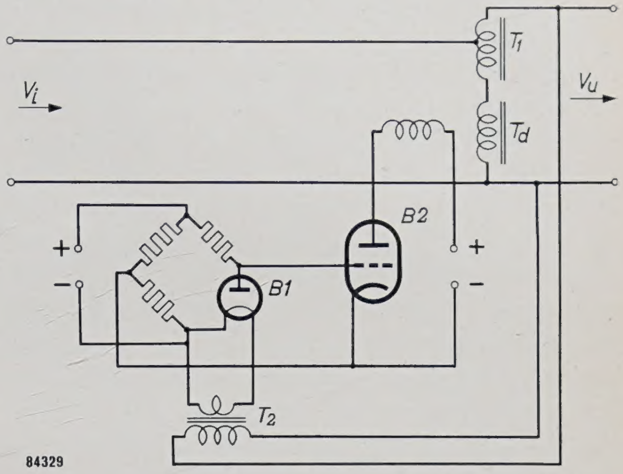

Fig. 10. Simplified circuit diagram of the equipments of fig. 9. The bridge incorporating the diode B_1 becomes unbalanced if the filament voltage of B_1 deviates from its nominal value; the bridge thus acts as detecting device.

Tube B_2 amplifies the error signal and controls the direct current through the transductor. An additional filter (not shown in fig. 10) suppresses the harmonics in the output voltage.

The stabilizers described here supply a more constant voltage than the "magnetic" stabilizers of fig. 6, as is evident from Table I. The speed of regulation, however, is somewhat lower. For many applications this is not important and this type of stabilizer is widely used.

A new fast-acting mains-voltage stabilizer

In the introduction, an application requiring a high speed of regulation was mentioned, viz. the supply of an X-ray tube used for making short-exposures, such as for diagnostic purposes. In this case, the high tension and the filament voltage of the X-ray tube must not deviate too much from their correct values: too high a voltage or too high an emission current would cause overloading and excessive blackening of the photographic material, while too low a voltage or current would cause insufficient blackening. The density of the photographic image is proportional to about the fifth power of the H.T. voltage, and to the eighth or tenth power of the filament voltage. The exposures are often very short (e.g. 0.02 sec.); hence a fast acting stabilizer is necessary.

The stabilizer to be described has been especially designed for those industrial and X-ray applications which require not only the regulation speed of the

magnetic stabilizers but also the frequency-independence and the accuracy of the feedback system. It is possible to combine these properties. To see how this may be done we need only consider what happens if a small, variable self-inductance is inserted between the points A, B in fig. 4 and 5. The introduction of this self-inductance will disturb the resonance. A system like this, however, as we have seen (fig. 2), will maintain a state that approximates closely to the state of resonance. The current through the saturated choke will rise, in order that the sum of the self-inductances in the resonant circuit remains substantially constant. Due to this, the voltage across the capacitor will also rise. The output voltage of the stabilizer can thus be controlled by means of an additional variable self-inductance. A certain output voltage will correspond to a given value of this self-inductance; this voltage is almost independent of the input voltage.

Here too, the variable self-inductance may be given the form of a transductor, energized by a feedback system similar to that shown in fig. 7. The result is that fluctuations of the output voltage, such as those caused by frequency fluctuations, are compensated by means of the transductor. It is true that this method of control is slow-acting, but the frequency variations concerned also occur slowly. The ordinary voltage fluctuations of the mains are rapidly compensated by the magnetic system.

This stabilizer also has a higher factor of improvement than that of the magnetic stabilizer alone. With regard to this better factor of improvement, the inertia of the feedback circuit does not constitute a serious drawback. This is due to the fact that the larger variations in the mains voltage (e.g. 10 V at a nominal mains-voltage of 220 V) do not as a rule occur suddenly, whilst the commoner smaller fluctuations are sufficiently stabilized by the magnetic stabilizer with its lower factor of improvement. Moreover, if the load or the load power factor is changed, the feedback will act so as to restore the output voltage to its correct value.

The detecting device

A convenient means of detecting small departures of an alternating voltage from its nominal r.m.s.-value is a thermistor bridge [2]), such as shown in *fig. 11.*

A typical voltage-current characteristic of a thermistor is shown in *fig. 12*, curve (*1*), whilst curve (*2*) shows the characteristic of a linear resistor. If these two elements are connected in series, the resultant characteristic is of the form shown in curve (*3*), which by suitable choice of resistor has a flat portion of the curve AB wherein a very small change in applied voltage will produce a very large

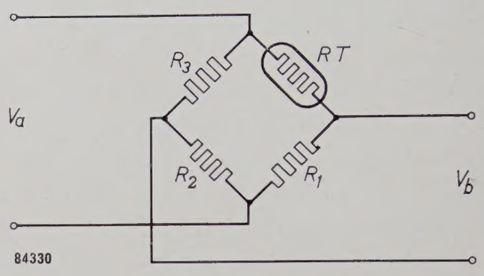

Fig. 11. Detecting device of the "Balham" stabilizer. V_a is proportional to the output voltage of the stabilizer. RT is a thermistor; its resistance is strongly dependent upon the temperature. At the normal value of V_a the bridge is just balanced. If V_a assumes a higher or lower value, the temperature of RT will change, so that the bridge becomes unbalanced. This causes a relationship similar to that of fig. 8 between V_a and V_b.

change in the circuit current, and consequently in the voltage developed across the linear resistor. This principle is employed in the bridge circuit of fig. 11; the resistors R_2 and R_3 are so chosen that the bridge is balanced if the system operates in the current range AB. Any small variation of the voltage supplied to the bridge (V_a) will then unbalance the bridge, and bring about a considerable variation of V_b. This applies not only to a direct voltage, but also to the r.m.s.-value of an alternating voltage, as the thermistor is incapable of following the rapid variations. The output voltage of the bridge will be either in phase or in antiphase with the supply voltage, according to whether the supply voltage is above or below the nominal value.

If a bridge of this type is used in a stabilizer, the peak shown in curve (*3*) of fig. 12 constitutes a difficulty, because of the fact that when the appara-

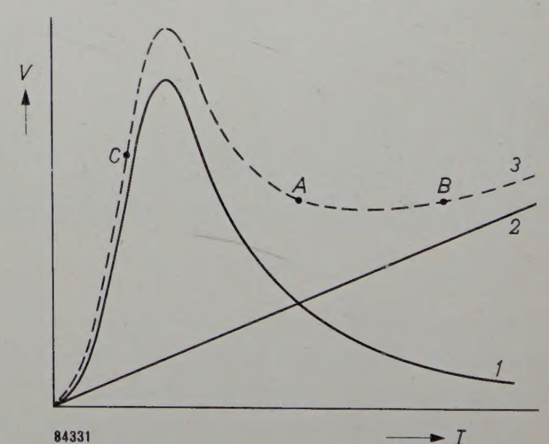

Fig. 12. Voltage-current characteristic of *1*) a thermistor, *2*) an ordinary resistor, *3*) a thermistor and resistor in series.

[2]) Philips Tech. Rev. 9, 239-248, 1947.

tus is switched on, the working point occurs somewhere in the vicinity of C. The bridge then produces a high output voltage which tends to increase the output voltage of the stabilizer, and thus to produce an upward shift of the working point. In this, however, it does not succeed, since the output voltage can only be varied by the transductor within certain limits. In order to raise the working point from C across the peak it would be necessary to raise the output voltage of the stabilizer (which feeds the bridge) for a short time above its nominal value, which, obviously, is not permissible.

In existing stabilizers with thermistor control special circuits with relays and electronic tubes are applied to overcome this obstacle [3]). In the apparatus under discussion the need for these special precautions has been obviated by the use of an *indirectly heated* thermistor, which is preheated by a spiral heating element surrounding it. The effect of this preheating is that the peak on the thermistor voltage-current characteristic is lowered considerably (curve *1*, *fig. 13*). If a resistor is

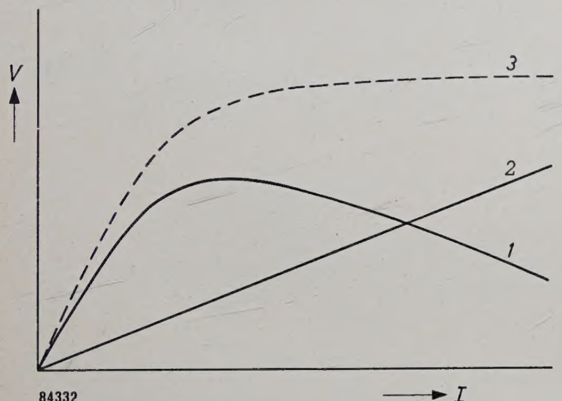

Fig. 13. The same as in fig. 12. Here, however, the thermistor is pre-heated by a separate heating element.

connected in series with such a thermistor the characteristic becomes as shown in curve (*3*). In this curve the long, flat portion necessary for high sensitivity has been retained, but as the maximum has disappeared, it is unnecessary to provide special circuits for moving the working point to a flat region.

The amplifier

If the output voltage of the stabilizer deviates from its nominal value, the thermistor bridge will produce a signal in phase or antiphase with the supply voltage. This signal must be used to alter the direct current through the transductor so as to restore the output voltage to its correct value. This is achieved by amplifying the bridge output voltage and applying it to the control grid of an output pentode (*fig. 14*) whose anode circuit incorporates the D.C. winding of the transductor *Td*. The screen

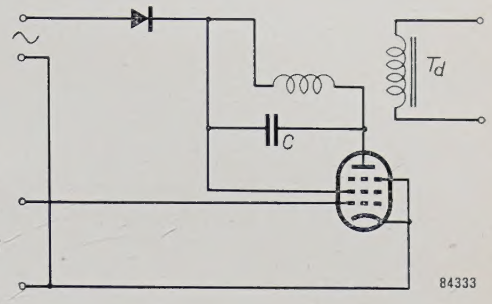

Fig. 14. Amplifier of the "Balham" stabilizer. A pulsating direct voltage is applied to anode and screen grid of the tube, so that this tube functions as a rectifier; the current through the D.C. winding of transductor *Td* in the anode circuit is dependent on the alternating voltage on the grid.

grid and anode of this tube are supplied with a rectified non-smoothed voltage in phase with the supply to the thermistor bridge. If the signal on the grid is in phase with the anode voltage, the mean anode current will increase by an amount dependent on the grid signal. Conversely, if the grid signal is in antiphase with the anode signal, the mean anode current will be reduced. The current flowing through the D.C. winding of the transductor is smoothed by the capacitor C.

This amplifier, which is at the same time a phase-sensitive detector, reacts to that component of the bridge output voltage that is in phase (or in antiphase) with the mains voltage. If the bridge is balanced for this component, there remains a low voltage with a phase difference of 90° with respect to the bridge voltage. This residual voltage is made up of components having the mains frequency and its third harmonic, and is produced due to the fact that the thermistor is not infinitely slow-acting, but changes its resistance perceptibly during one period of the mains frequency [4]). As the two components lead the mains voltage by 90°, they do not influence the current through the transductor. The filters used in previous thermistor circuits are, therefore, superfluous.

The complete circuit

The complete circuit of the detecting and regulating devices is given in *fig. 15*.

The transformer T_1, which is connected to the output voltage of the stabilizer, supplies the anode and filament voltage for the amplifying tube B_1, and in addition it also supplies the thermistor heater and the thermistor bridge. The latter consists of the potentiometer R_2, the resistor R_3 and the thermistor RT_2. The output voltage of the bridge is coupled to

[3]) G. N. Patchett, Precision A. C. voltage stabilizers, Part VI, Electronic Eng. 22, 499-503, 1950.

[4]) See the article referred to in [3]), Part I, pp. 374-377.

the control grid of an amplifying valve B_1. The anode supply of this valve is derived from the unsmoothed supply to the phase-sensitive detector via the smoothing network R_4-R_5-C_1-C_2.

Since the thermistor is a temperature-sensitive element, variations in ambient temperature would normally tend to cause drifts in the stabilizer output voltage. To compensate for such effects, the thermistor heater is shunted by another thermistor RT_1 of such a size that self-heating due to its own current flow is small. Increasing the ambient temperature produces a decrease in the resistance of RT_1 and a corresponding reduction in the heater current of RT_2. This compensation is sufficiently accurate in the temperature range of 10-50 °C.

In the event of failure of the anode current of B_1 for any reason, the self-inductance of the transductor and consequently the output voltage would rise to a high value. To prevent this, a relay RL_1 has been incorporated in the anode circuit of the tube. A contact RL_1c on the relay short-circuits the A.C.-winding of the transductor if the relay becomes de-energized. Since the short-circuiting of the transductor causes a low output voltage of the stabilizer, the load circuit is protected against excessive voltage in the event of tube failure.

The relay also facilitates the switching on of the circuit in the following way. When the stabilizer is connected to the mains, B_2 at first does not pass anode current due to its cold filament. Relay RL_1 is therefore not energized. Under these conditions thermistor RT_2 is provided with an excess heater voltage via winding S_3 and contact RL_1a, and is *rapidly* brought to its correct operating temperature. When the tube current reaches a value sufficient to close the relay, the normal heater voltage is applied to the RT_2 heater via contact RL_1b, and R_1 which is adjusted so that the current has the correct value.

The current at which the relay closes is adjusted by the parallel resistor R_8 to a value which does not produce a surge of stabilizer output voltage when contact RL_1c across the transductor is opened. Contact RL_1d disconnects R_8 when the relay is energized, enabling the latter to remain closed down to the lower safety limit of the tube current.

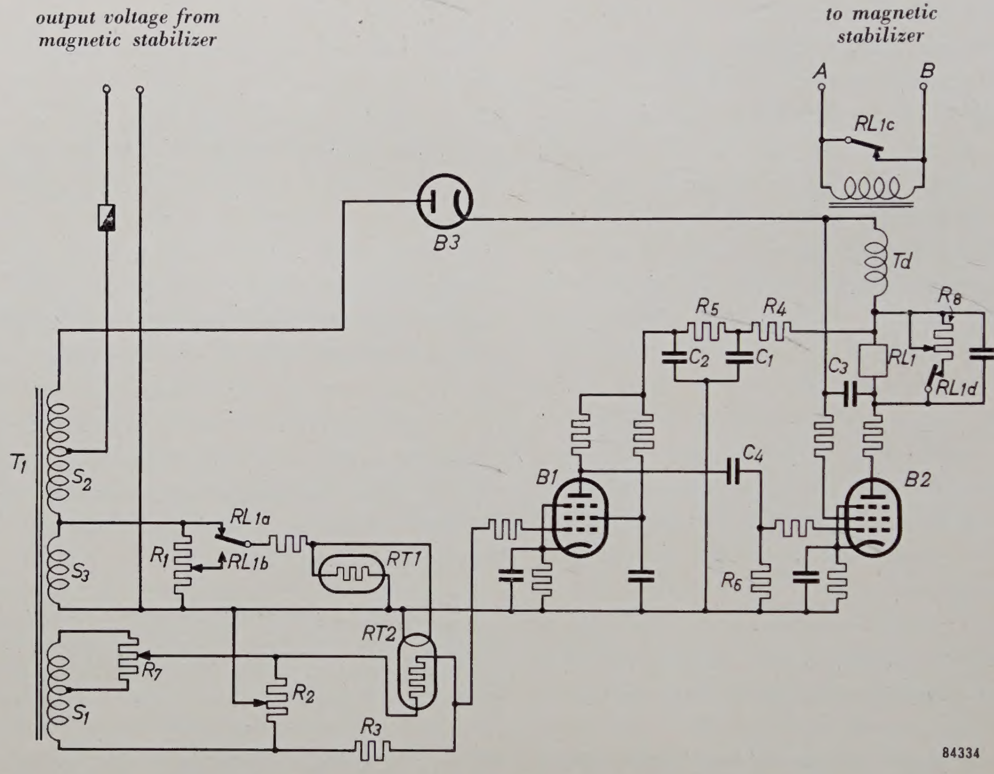

Fig. 15. Circuit diagram of the "Balham" stabilizer. The "magnetic" part is not shown. The A.C. coil of the transductor is assumed to be connected between the points A and B of fig. 4 or fig. 5.

The anode supply of B_1 is derived from the junction of RL_1 and the D.C. winding of the transductor. Hence a fraction of the anode voltage of B_2 is fed back to the grid via C_4, resulting in a negative feedback at frequencies within the passband of the network C_4R_6. This feedback, which is analogous to "velocity feedback" in a servo system, prevents low frequency oscillations in the loop of fig. 7. The presence of C_4 prevents D.C. feedback and thus the "improvement factor" is not reduced.

The inductance of the D.C. winding of the transductor, together with the parallel condenser C_3 and the internal resistance of B_2, causes a phase shift for sinusoidally varying signals which becomes larger at increasing frequency. The

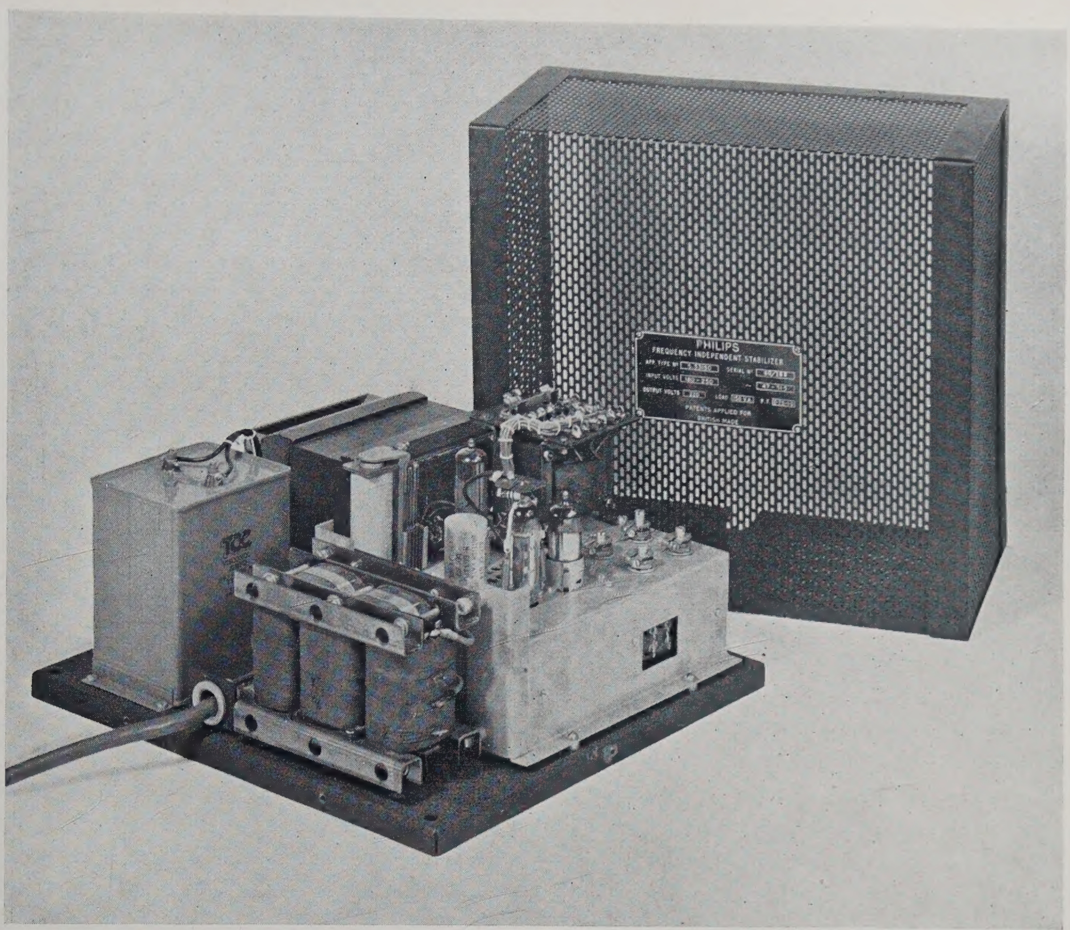

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Fig. 16. Stabilizer for 150 VA output based on the circuit of fig. 15.

inertia of the thermistor produces a similar effect. For the frequency at which the total phase shift in the loop of fig. 7 is 180°, the total loop amplification should be smaller than unity to prevent oscillation. The above-mentioned frequency-dependent feedback via C_4 will counteract the phase shift due to C_3 and the transductor inductance, except at frequencies so high as to be attenuated sufficiently by the inertia of the thermistor.

The output voltage of the stabilizer can be adjusted within certain limits by means of R_7. The position of this potentiometer determines the value of the primary voltage of T_1 at which the bridge is balanced.

The complete voltage stabilizer is shown in *fig. 16*. The maximum load is 150 VA.

The stabilizer will compensate for any of the following changes:

Input voltage changes from 180 to 250 V.

Frequency changes in the range 47-51.5 cps.

Load changes in the range 0-150 VA.

Power Factor changes in the range 0.75-1.0.

Temperature changes from 10 °C to 50 °C.

For any combination of operating conditions within the above ranges, the output voltage will lie within ± 0.5 % of the nominal voltage.

For more limited changes the deviation will be less (see table I).

Summary. After surveying the requirements which mains voltage stabilizers should meet for various applications, this article gives a brief description of "magnetic" stabilizers and "feedback" stabilizers. The former operate on the basis of the non-linear relationship between magnetic field strength and induction in the soft-iron core of a choke. Although these stabilizers are very fast-acting, their applications are restricted by the fact that the output voltage is dependent on the mains frequency and the load power factor.

The feedback type of stabilizer, an example of which is also given, has the drawback in some applications of being rather more slowly acting, due to the fact that the voltage-sensitive element, incorporated in the feedback circuit is necessarily slow-acting. This type, however, achieves a high precision and frequency-independence. A new type of stabilizer is described, developed especially for an application requiring fast operation. It combines the favourable features of both above-mentioned types, because here the output voltage of an ordinary magnetic stabilizer is kept constant with the aid of a simple feedback system. The voltage-sensitive element is a bridge circuit incorporating a thermistor. Indirect heating of the latter prevents difficulties during warming up. The influence of the ambient temperature is compensated by a second thermistor. A stability to within ± 0.5 % over the temperature range 10-50 °C can be achieved for mains-voltage fluctuations between 180 V and 250 V, frequency variations between 47 c/s and 51.5 c/s, and load variations from no load to full load. The recovery time amounts to between one and two cycles of the mains frequency.

SUPPLY UNITS FOR AIRFIELD LIGHT BEACONING SYSTEMS

by Th. HEHENKAMP.

621.326.062:628.971.8:629.139.1

Resonance phenomena are essential features of high-frequency technique, but rarely find application in power engineering. Nevertheless, a resonant circuit occasionally offers the best solution to a particular power problem, for example that of supplying a constant current to lamps in series. The principle of resonant circuits — which are linked to such names as Boucherot and Steinmetz — is now more than 60 years old; it does not often happen in electrical engineering that a principle so long known and so much used in some fields has yet been neglected in others.

To enable aircraft to land safely, in darkness or in mist, it is necessary to equip the airfields with effective light beacons. An article describing in detail the system of beaconing employed on modern airfields appeared in a recent issue of this Review [1]). As explained there, the beacon lights are fitted with low-voltage incandescent lamps (6 to 30 V), since this permits the use of a compact filament, whereby a more concentrated light beam can be attained.

A direct supply to these lamps in parallel is impracticable, since it would necessitate unduly thick cables. On the other hand, supplying the lamp direct in series, although possible, has the disadvantage that breakdown of one lamp would interrupt the entire chain [2]). Accordingly, every lamp in the system, whether supplied in series or in parallel, is provided with a separate transformer which matches it to the appropriate mains voltage or current. Such lamp transformers are made for different powers and in various types, according to the particular method of installation (buried, incorporated in the light itself, etc.). The transformers are fed from a special supply unit equipped with magnetic switches, enabling the required runway lights and their luminous intensities to be selected. Such supply units are housed in transformer stations located as close to the runways as possible and often below ground level. They can be operated either in the station itself or from the control tower, which is then equipped with a desk incorporating the necessary controls and monitor lamps.

Before discussing the details of this equipment, let us consider whether it is better to connect the primary coils of the individual lamp transformers in parallel, or in series.

[1]) J. B. de Boer, Philips tech. Rev. **16**, 273-286, 1954/55 (No. 10).

[2]) This can be prevented by connecting a special cartridge fuse in parallel with each lamp, a method sometimes employed in street lighting systems (see Philips tech. Rev. **6**, 105-109, 1941); however, it is not considered reliable enough for airfield lighting.

Parallel or series supply?

Of all the various considerations upon which the choice of the supply system should be based, the most important are capital expenditure and reliability.

Capital expenditure

To minimize capital expenditure as far as possible, it is necessary to employ thin cables. Consequently, an appreciable voltage drop must be tolerated; in general, the associated loss of energy is not important, since the costs arising from it are small as compared with the cable depreciation (one runway 2 miles long may require cable costing £ 5000).

With a parallel supply system, the toleration of a substantial voltage drop gives rise to a noticeable difference in luminous intensity between the lamps at opposite ends of the cable. In a series system, on the other hand, such a drop produces no difference in luminous intensity, since the current is the same in all the lamps. Although this is a very important advantage of the series system, it is not in itself reason enough to conclude that such a system is best for the present purpose. Before reaching such a conclusion, we must compare the two systems more precisely by determining the relative voltage drop in them when the cable cost, or, to simplify matters, the volume of copper (by far the biggest contribution to the cost of the cable), is the same in both (for a given copper volume, the number of cores has little effect on the cost).

Let us consider the case of a runway (length l), flanked on both sides by n lights each of nominal power P arranged at regular intervals; in practice, n is usually between 50 and 100. With a parallel supply system (*fig. 1a*), the voltage across the first lamp is E_p (suffix $_\mathrm{p}$ for parallel) and that across the last $E_\mathrm{p} - \Delta E_\mathrm{p}$. Owing to the positive temperature coefficient of resistance of the tungsten filaments, the difference in current between the individual lamps will be appreciably smaller than the difference

in voltage. To simplify the following calculation, let us ignore current differences, that is, assume that the current supplied to each lamp transformer is $I_p = P/E_p$.

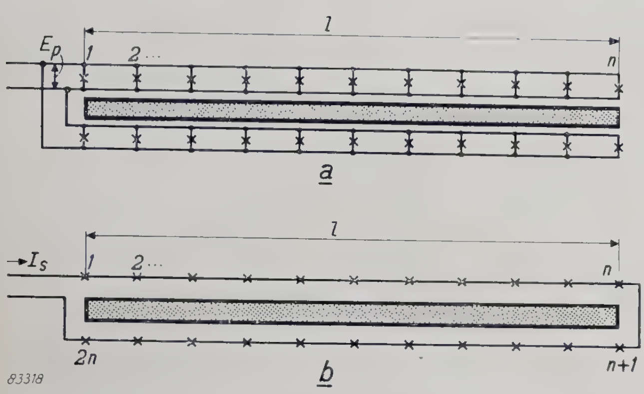

Fig. 1. Runway with n lights on each side. *a*) Parallel supply, *b*) series supply.

We then find that the potential difference ΔE_p is:

$$\Delta E_p = \frac{n \varrho l P}{A_p E_p},$$

where A_p is the cross-sectional area of one cable core, and ϱ the specific resistance of copper. Accordingly, the relative potential difference is:

$$e_p = \frac{\Delta E_p}{E_p} = \frac{n \varrho l P}{A_p {E_p}^2}. \quad \ldots \quad (1)$$

With a similar system of lights supplied in series (fig. 1*b*), with a current I_s, the voltage drop in the cable will be:

$$\Delta E_s = \frac{2 \varrho l I_s}{A_s} \text{ (suffix }_s\text{ for series).}$$

The total voltage across the series system is then:

$$E_s = \frac{2nP}{I_s} + \Delta E_s .$$

Again introducing the relative voltage drop ($e_s = \Delta E_s/E_s$), we have:

$$e_s\,(1 - e_s) = \frac{4 n \varrho l P}{A_s {E_s}^2}. \quad \ldots \quad (2)$$

Since in a parallel supply system the runway is flanked on each side by two cores of cross-sectional area A_p, and in a series system by one core of area A_s on each side, the condition of equal copper volume is satisfied if $2A_p = A_s$.

From (1) and (2), we have, therefore:

$$e_p {E_p}^2 = \tfrac{1}{2} e_s\,(1 - e_s) {E_s}^2.$$

Assuming that the supply voltages of the two systems are the same, the relationship between e_p and e_s becomes:

$$e_p = \tfrac{1}{2} e_s\,(1 - e_s). \quad \ldots\ldots \quad (3)$$

This relationship is shown graphically in *fig. 2*.

Because in a parallel system the difference in voltage gives rise to a difference in luminous intensity between the lamps, e_p is limited, in the absence of special measures, to about 3% (the fact that the lamps themselves, the optical systems and their adjustment, exhibit a certain amount of spread is taken into account).

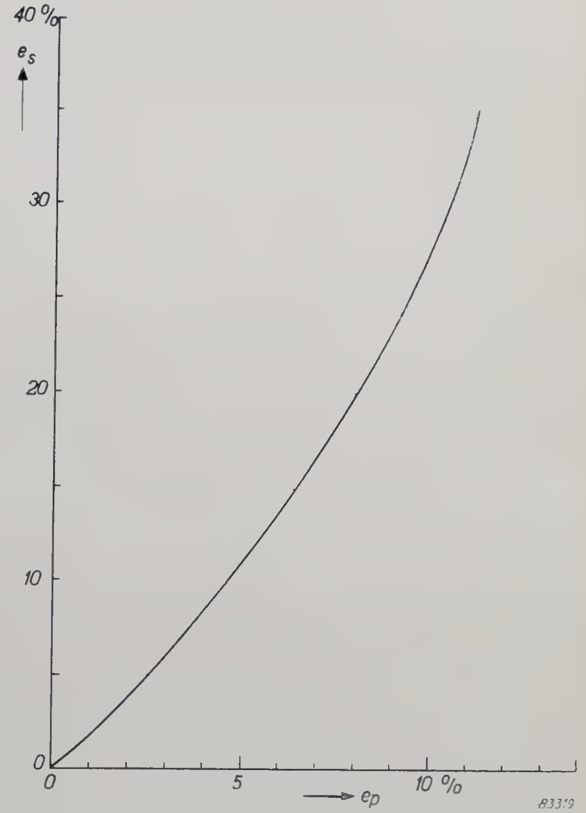

Fig. 2. Relationship between the relative voltage losses e_s and e_p, in a series and a parallel system respectively, the copper volume, mains voltage, and overall runway lighting power being the same in both cases.

According to fig. 2, $e_p = 3\%$ corresponds to $e_s = 6\%$ in a series system with the same copper volume and supply voltage. However, in a series system e_s may be taken very much higher (e.g. 25%), and a very considerable saving in cable costs thus procured without impairing the lighting, provided that the particular supply unit is designed for a correspondingly higher power level; e_s is then limited only by the last-mentioned condition, usually to about 25 or 30%.

To compensate for the fact that in a parallel system the voltage across the more remote lamps is lower than that across the nearer ones, the lamp transformers may be provided with suitable tappings. This enables the limit on e_p to be raised to

about 10%. (If it were any higher, the lamps would show too perceptible a difference in luminous intensity when operated at reduced power, owing to the fact that the resistance of the lamps is then so low as to increase the relative effect of the cable resistance considerably. The lamps are normally operated at reduced power to avoid glare, the full light yield being required only in foggy conditions by day). From fig. 2, e_p = 10 % corresponds to e_s = 28 % in a series system. If this is taken as the maximum limit, no further saving in cable costs will be possible; nevertheless, the series system is better in that it ensures a uniform light output, whereas in a parallel system with e_p = 10 % the irregularity of the lighting is only just within the limits of tolerance.

In many cases there are still stronger objections to the use of a parallel circuit. A high voltage supply is less acceptable in such a system than in a series circuit where there is never a high potential difference between adjacent points (except, of course, inside the transformer station). This makes problems of insulation and protection far easier to solve than in a parallel system.

Reliability

From the point of view of reliability, it is extremely important that the effects of minor local defects and breakdowns be reduced as far as possible. The transformers, and the actual lights, are often placed below, or only just above, ground level, so that various clips and pressure contacts (e.g. those of the lamp holders) are exposed to a corrosive atmosphere. Accordingly, the possibility of some contact resistance must always be taken into consideration. One of the advantages of series circuits whose current supply does not depend upon the load resistance is that contact resistances do not affect the current in the circuit, but merely cause a slight voltage drop across the bad contact. Moreover, a contact broken electrically but mechanically intact (e.g. that between the lamp cap and the lamp holder) may easily be restored by the high voltage applied to it when the installation is switched on. With a parallel supply, on the other hand, contact resistances present a far more difficult problem, since they may cause under-loading of the particular lamp or group of lamps, in some cases so much so as to extinguish it completely.

Interference of another kind is caused by short-circuits, as will now be explained. The lights may be damaged mechanically by aircraft or other traffic, or in the course of work carried out along the runway, in such a way as to cause a short-circuit. In a series system, the worst that can then happen is failure of the lights affected, the others continuing to operate normally. With parallel feeding, however, the possibility of shorting necessitates expensive safety measures; for example, each light must be provided with a fuse in series, and the positioning of such fuses usually presents an awkward problem. Moreover, it is difficult to ensure that they will function reliably, since, owing to the relatively high cable resistance tolerated for reasons of economy, the short-circuit current only slightly exceeds the normal operating current. Taking into consideration also the fact that it is simpler to standardise the supply units and transformers of a series system, and to compensate for changes taking place during or after the installation of the system, it will be evident that in general a series supply is best suited to the requirements of airfield beaconing systems.

However, approach lights may be an exception to this general rule, particularly high power lights. In approach lighting, a large number of lights, consuming a great deal of power, are concentrated in a limited zone some distance from the runway, so that the risk of damage to them or to the cables is relatively small. These lights and the associated transformers may therefore be raised quite a long way above ground level; hence the operating conditions of this part of the installation are more favourable than those of the actual runway lights. Accordingly, the lights are less likely to become defective and are easier to inspect and maintain. Moreover, the fitting of the fuses required in a parellel system is relatively simpler. Here, three-phase mains, with the lights distributed evenly between the three phases, is often the best solution.

Boucherot's constant-current bridge circuit

To employ a series system to full advantage, it is essential to make the supply current independent of the load. This can be accomplished by means of a transformer with a movable secondary coil, a method already used in street lighting for several decades. However, another method, involving no moving parts, has become more popular during the past 15 years. It is based on a bridge circuit originally designed by Boucherot (1890) and developed by Steinmetz [3]).

In the ideal case (exact resonance and no losses), this circuit exhibits two special characteristics:

1) the output current does not depend upon the size and nature of the load impedance, and
2) with resistive loading, the input current is in phase with the supply voltage from the mains.

We are most concerned with the first characteristic, since the second merely indicates that with resistive loading the power factor is unity.

Theory of the constant current bridge (neglecting losses)

To demonstrate the above characteristics, let us now consider a bridge circuit with impedance Z_1 in

[3]) See, for example, R. R. Miner, Resonant-type constant-current regulators, Trans. Amer. Inst. El. Engrs. **58**, 822-829, 1939.

one pair of opposite arms, impedance Z_2 in the other pair of arms (*fig. 3a*), impedance Z in the diagonal link and a supply voltage E_0 across the other diagonal.

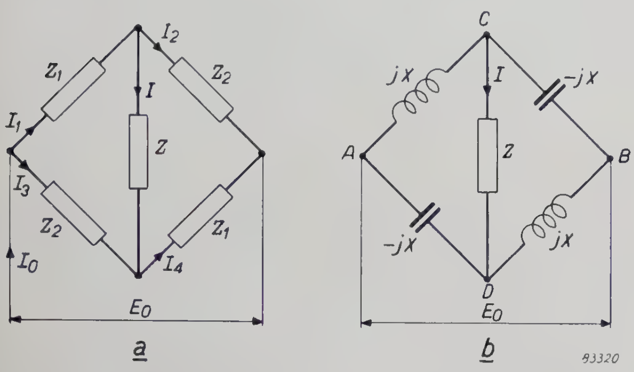

Fig. 3. *a*) Bridge circuit with two impedances Z_1 and two impedances Z_2 in the arms, and a load impedance Z in one of the diagonals.
b) Putting $Z_1 = +jX$ and $Z_2 = -jX$ in *a*), we have a Boucherot bridge, in which the current I in Z is independent of Z.

From the equations (for notation see fig. 3*a*)

$$I = I_1 - I_2$$

and

$$E_0 = I_1 Z_1 + I_2 Z_2,$$

after eliminating I_1, we have:

$$I = \frac{E_0 - I_2(Z_1 + Z_2)}{Z_1}. \quad \dots \quad (4)$$

We now choose Z_1 and Z_2 such that

$$Z_1 = -Z_2 \quad \dots \dots \quad (5)$$

(implying that the two are in resonance at the mains frequency).

From (4), we then get:

$$I = \frac{E_0}{Z_1}, \quad \dots \dots \quad (6)$$

showing that the output current I is in fact independent of the load impedance Z.

We find that the current from the mains (I_0) is:

$$I_0 = I_1 + I_3,$$

and that:

$$I_1 Z_1 + IZ - I_3 Z_2 = 0.$$

Eliminating I_1 from these two equations, we have:

$$I_0 = \frac{I_3(Z_1 + Z_2) - IZ}{Z_1},$$

which becomes with the aid of formulae (5) and (6):

$$I_0 = -\frac{Z}{Z_1^2} E_0. \quad \dots \dots \quad (7)$$

To satisfy the resonance condition (5) fully, Z_1 and Z_2 must be entirely devoid of losses, that is, pure reactances: $Z_1 = jX$ and $Z_2 = -jX$ (fig. 3*b*). Accordingly, the denominator in formula (7), Z_1^2, is real: hence it is seen that the input current will be in phase with the supply voltage (E_0) if the load impedance (Z) is a pure resistance. This, then, demonstrates the second characteristic.

The circuit considered has yet another unusual feature (this time an undesirable one), namely that the currents in the four arms, I_1, I_2, I_3 and I_4, can not be individually determined; application of Kirchhoff's laws to the junctions and loops of the bridge produce a set of interdependent equations. However, the sums of the currents in opposite arms are determinate:

$$I_1 + I_4 = I\left(1 - \frac{Z}{Z_1}\right) \quad \dots \dots \quad (8)$$

and

$$I_2 + I_3 = -I\left(1 - \frac{Z}{Z_1}\right). \quad \dots \dots \quad (9)$$

Provided that formulae (8) and (9) hold good, the currents in the individual arms may assume any value. This will be evident from the fact that the impedance of circuit *ACBDA* (fig. 3*b*) is zero, so that any current can flow in this circuit without affecting the voltages across the diagonals.

Another point to bear in mind in this connection is that the individual currents referred to are indeterminate only at resonance and with circuit elements devoid of losses, as will be seen from the following. In general:

$$I_1 + I_3 = I_2 + I_4 \quad \dots \quad (10)$$

and

$$I_1 Z_1 + I_2 Z_2 = I_3 Z_2 + I_4 Z_1.$$

Hence we have:

$$(I_1 - I_4) Z_1 = -(I_1 - I_4) Z_2. \quad \dots \quad (11)$$

Now, if $Z_1 \neq -Z_2$ (owing to a resistive term in Z_1 and Z_2, or a deviation from resonance, or a combination of the two), it follows from formulae (11) and (10) that:

$$I_1 = I_4 \text{ and } I_2 = I_3 \quad \dots \dots \quad (12)$$

It will be seen, then, that the individual arm currents associated with any variation from formula (5) are determinate, these currents being indeterminate only when formula (5) holds; at the same time, this formula must hold good to give the circuit the required constant current characteristic, (equation 6). In practice, this contradictory state of affairs necessitates a compromise, as will now be explained.

Theory of the bridge taking losses into account

The practical bridge circuit differs from the purely theoretical one so far considered, in that the circuit elements are not entirely devoid of losses. The capacitor losses are usually insignificant, but for economic reasons the losses in the coils cannot be reduced till their effect is made negligible. Hence it is necessary to ascertain the precise nature of this effect.

With circuit elements not entirely devoid of losses, condition (5) cannot be fully satisfied, and, as we have already seen, failure to satisfy this condition would eliminate the indeterminacy of the currents. However, practical experience has shown that the currents may still be unduly dependent on circumstances: one capacitor voltage may be, say, 5 times the other. This is, of course, highly undesirable, since it would necessitate over-dimensioning the capacitors and coils to a very considerable extent.

This can be avoided in several ways. From formula (12), assuming that formula (5) does not hold good, the currents in the coils are theoretically equal and 180° out of phase, and it is only by chance that in reality these currents usually differ.

Identical currents can be procured by placing the two coils on a common iron core (*fig. 4a*). A variant of this method is to couple the coils by means of a 1 : 1 transformer, thus maintaining the capacitor voltages the same (fig. 4*b*).

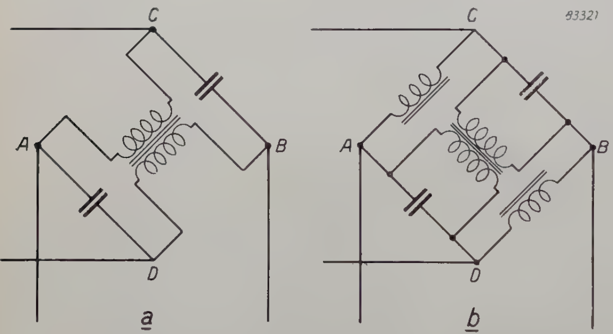

Fig. 4. The indeterminacy of the four arm currents in the Boucherot bridge can be eliminated by fitting the two coils on one core (*a*), or by coupling the capacitors across a 1 : 1 transformer (*b*).

Another method — employed in Philips equipment — involves a deliberate deviation from condition (5), by detuning slightly, apart from the inevitable deviation caused by the losses. Not only does this method enable the indeterminacy of the arm currents, which persists despite losses, to be avoided, but it also precludes all possibility that a simultaneous failure of several lamps will cause the remainder to be overloaded (a possibility to which we shall refer again in the section on lamp transformers). The permissible degree of detuning depends of course, on the extent to which it detracts from the constancy of the output current. In view of this, the calculation of the output current will now be repeated, on the assumption that neither aspect of condition (5) is fulfilled.

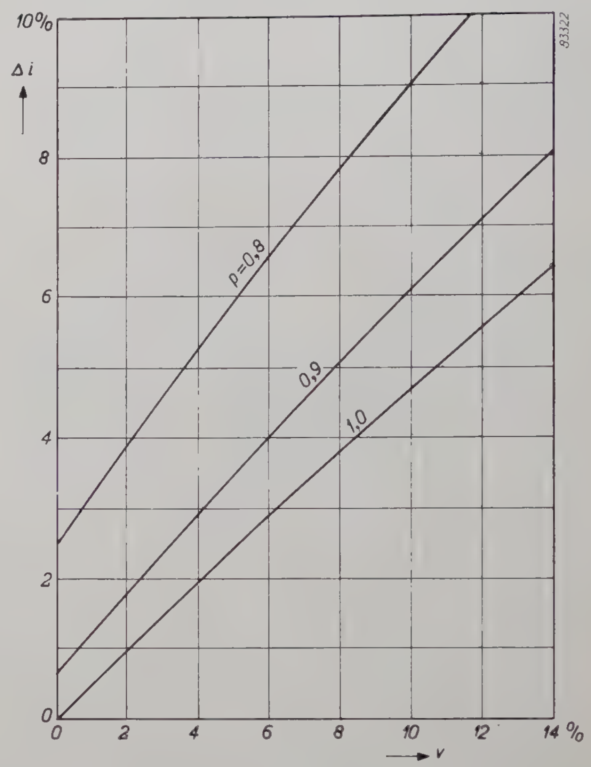

Fig. 5. The relative current difference Δi plotted against the relative loss v in the Boucherot bridge, for various values of the detuning p. Almost identical curves hold good for the reciprocal values of p.

The general equation defining the output I is:

$$I = \frac{Z_2 - Z_1}{ZZ_1 + ZZ_2 + 2Z_1Z_2} E_0 \, . \quad . \quad . \quad (13)$$

The load impedance Z, formed by the series system of lamp transformers each loaded with an incandescent lamp [4]), is for all practical purposes a pure resistance (R). Now, if Z_2 is made up of a coil with self-inductance L and a resistance r, and Z_1 of a capacitor C, assumed to be free of losses, and taking ω as the angular frequency of the mains, we have:

$$Z = R, \qquad Z_1 = \frac{1}{j\omega C}, \qquad Z_2 = r + j\omega L.$$

Next, we introduce two parameters, the one, $v = r/R$, referring to the losses, and the other, $p = \omega/\omega_0$ (where $\omega_0{}^2 = 1/LC$), to the detuning, so that (13)

[4]) The case of unloaded transformers (defective lamps) will be discussed later.

may be written:

$$I = \frac{E_0}{R} f(p,v).$$

The current reaches a maximum, I_{max}, in the event of a short-circuit ($R = 0$, $v = \infty$). The relative current difference, $\Delta i = (I_{max} - I)/I$, is here adopted as a measure of the deviation from constant current output.

Fig. 5 shows Δi plotted against v for various values of p. It is seen from this diagram that the increase in Δi produced by detuning from $p = 1$ (resonance) to $p = 0.9$ or 1.1 will be small, provided that v is not unduly large. For $v = 6\%$, for example, Δi will increase from 3% to 4%, but any further detuning will cause a sharp rise in Δi.

Now, detuning to $p = 0.9$ is enough to reduce the indeterminacy of the arm currents appreciably, particularly if combined with one of the methods illustrated in fig. 4.

Description of a practical supply unit

On the basis of the theoretical arguments so far considered, Philips have designed a series of supply units ranging from 1.5 to 25 kVA. *Fig. 6* is a simplified circuit diagram. Here, *ACBDA* is Boucherot's bridge; the input diagonal. *AB*, is connected to

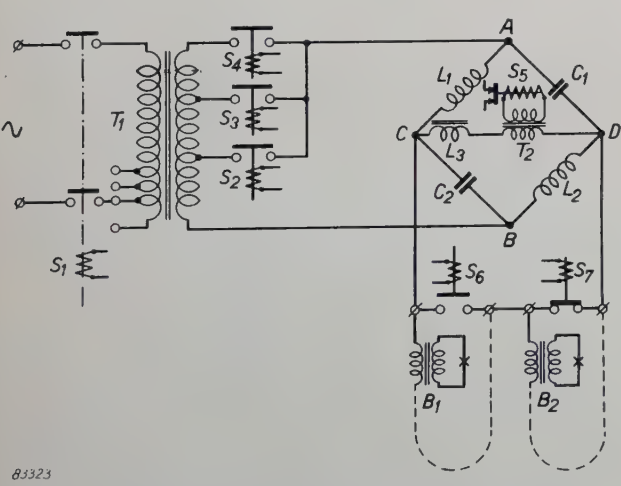

Fig. 6. Simplified circuit diagram of a low power supply unit. S_1 main switch; T_1 isolating transformer; S_2, S_3, S_4 luminous intensity selectors (varying in number from 0 to 5); *ABCD* Boucherot bridge; L_3 choke to limit the output voltage when the runway circuit is interrupted; T_2 current transformer; S_5 relay, which, when energised, opens mains switch S_1; B_1, B_2 runway light circuits; S_6, S_7 runway selectors.

the H.T. winding of a transformer, T_1, whose primary is supplied from the mains. The primary coil is provided with tappings, to enable adjustments to be made according to the average local mains voltage (it is seen from formula (6) that I varies in proportion to the supply voltage (E_0) of the bridge).

It is usual to make the luminous intensity of the installation variable in a number of steps; T_1 must therefore also be provided with tappings on the secondary. The bridge is connected at will to any of the secondary tappings by means of magnetic switches — so-called "luminous intensity selectors" — S_2, S_3 and S_4. It will be evident that only one selector may be energised at any given moment, since otherwise transformer T_1 would be short-circuited; hence the selectors are interlocked.

The output terminals of the bridge, C and D, may be connected to the circuit of runway lights either direct, that is, if the voltage across the series circuit is less than 1500 V, or via a transformer (not shown in the diagram), if a higher voltage is employed. It is often possible to operate two or more circuits alternately with one supply unit to light, say, the two directions of a single runway, or two alternative runways. These circuits are then connected in series and shunted separately by magnetic switches, known as runway selectors (fig. 6 shows two circuits and therefore includes two runway selectors, S_6 and S_7). On runways the circuits must operate one at a time; hence the runway selectors must also be interlocked. On taxiways, on the other hand, the lights must often be able to operate simultaneously in various combinations.

Interlock system

Various methods of interlocking magnetic switches are known; most of them involve auxiliary contacts designed to interrupt the connections to the switch energising coils. Where a large number of switches are to be interlocked, the number of auxiliary contacts required will be very large, and the risk of breakdowns therefore considerable.

In view of this risk, we have developed a system in which a high degree of reliability in service is ensured by minimizing the number of auxiliary contacts. It is based on the principle that the impedance of the energising coil of a magnetic switch is very much higher when the switch is closed than when it is open. It will be seen from the diagram (*fig. 7*) that one end of each energising coil is connected to a particular point P, which, in turn, is connected, via a capacitor (C_1) and a coil (L) in series, to one pole (*1*) of the mains supplying the control voltage. The other end of each coil can be connected to pole *2* of the mains via the push button (D_1, D_2, ...) appropriate to the particular switch. The ordinary change-over contacts, each in parallel with the associated push button, are indicated by *OC*, and the auxiliary contacts, to shunt capacitor C_1 across a low resistance R (almost a short-circuit), by *CC*. The series circuit C_1-L is resonant at the mains frequency; hence its impedance is low.

Now, let us assume that button D_1 is pressed. The coil of switch S_1 is then energised normally, so that it closes the main contact of S_1 (not shown in the diagram), and transfer contact *OC* and auxiliary contact *CC*. With the closing of *CC*, capacitor C_1 is "shorted" across resistance R (acting as a damping resistor to suppress the contact spark). For practical purposes, only coil L is then left between points P and *1*. (Owing to the

voltage drop in L, the voltage across the coil of S_1 would then be about 10% lower than the control voltage. In order to prevent S_1 from opening, even in the most unfavourable circumstances, a special capacitor C_2 is provided which, together with the coil of S_1, forms an impedance higher than that of the coil alone.)

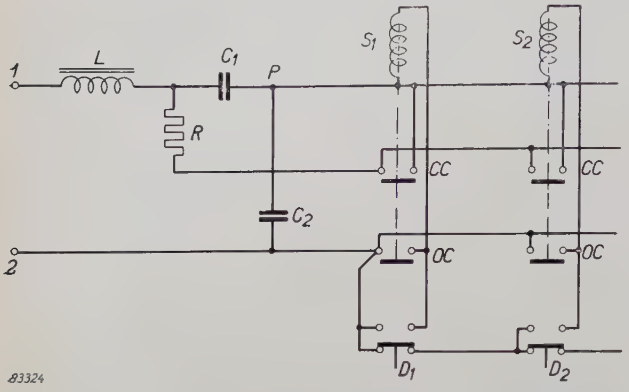

Fig. 7. Interlock mechanism to preclude the possibility that two or more switches (with energising coils S_1, S_2 ......) will be closed at the same time. *1*, *2* terminals of the mains supplying the control voltage; L-C_1 resonant circuit; R damping resistor; CC short-circuit contacts of C_1; OC transfer contacts C_2 auxiliary capacitor; D_1, D_2 ...... push buttons.

Next, button D_2 is pressed. This brings the coil of S_2 in parallel with that of S_1, and the two coils together in series with L. Since switch S_2 is still open, the impedance of its coil is low, and most of the control voltage is therefore dropped across L; hence the coil of S_2 is not sufficiently energised to close this switch, and the energising current of S_1 is so weakened that it opens. However, the shunt across C_1 is then interrupted; the impedance of circuit L-C_1 decreases and the energising current in S_2 becomes strong enough to close it.

Any number of switches may be employed, but it is impossible to energise more than one of them at a time. The pressing of any one of the buttons opens the switch already closed, and closes the one controlled by the particular button. As will be seen from the oscillogram shown in *fig. 8*, there is an interval of 3½ cycles between the opening of the one switch and the closing of the other.

To preclude all possibility that two or more switches will be closed at the same time (owing to two or more buttons being pressed simultaneously), the push buttons are connected in series in the usual way.

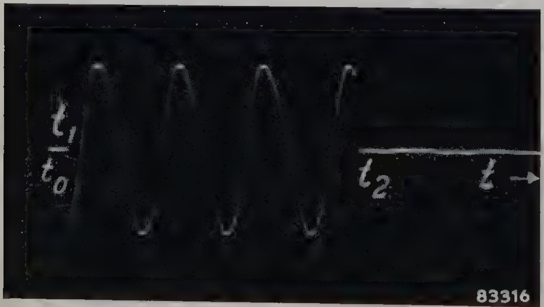

Fig. 8. Oscillogram referring to fig. 7. At $t = t_0$, the push button of S_2 is pressed, with S_1 already closed. After a certain delay, depending upon the design of S_1, the contacts of S_1 open, that is at $t = t_1$. A 50 c/s voltage then appears on the oscilloscope; this voltage is suppressed when S_2 closes (at $t = t_2$). The interval between t_1 and t_2 is 3½ cycles.

A closer analysis shows that it is advisable to detune the L-C_1 circuit slightly with respect to the mains frequency, since this reduces the adverse effect of the control cable resistance considerably. In fact, it enables this effect to be made much smaller than in ordinary interlock circuits, so that it is possible to employ relatively longer and/or thinner cables for the control current.

Limiting the output voltage

A break in the output circuit, say, owing to the cable being damaged, would cause a considerable increase in voltage. To avoid this, an auxiliary circuit comprising a choke (L_3) and a current transformer (T_2) is connected between output terminals C and D (fig. 6). Coil L_3 is so constructed that the current produced in it by a normal voltage between C and is negligible. However, if the voltage increases, owing to a break in the external circuit, the core of L_3 becomes saturated, thus limiting the voltage to less than twice its nominal value (*fig. 9*), despite the fact that the auxiliary circuit then carries the full current.

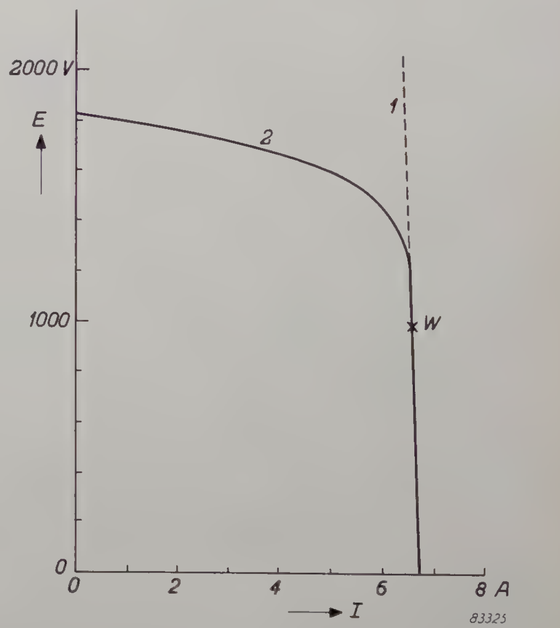

Fig. 9. The output voltage A plotted against the current in the external circuit I, (*1*) without limiter L_3-T_2 (fig. 6), (*2*) with the limiter in circuit. W full-load working point.

This current then energises a relay (S_5) via a current transformer. The break contact of this relay, in series with the energising coil of mains switch S_1 (fig. 6), therefore opens; thus, any break in the output circuit isolates the entire supply unit from the mains.

Mechanical design

All the components of the supply unit are accommodated in a single cabinet. At the front of this

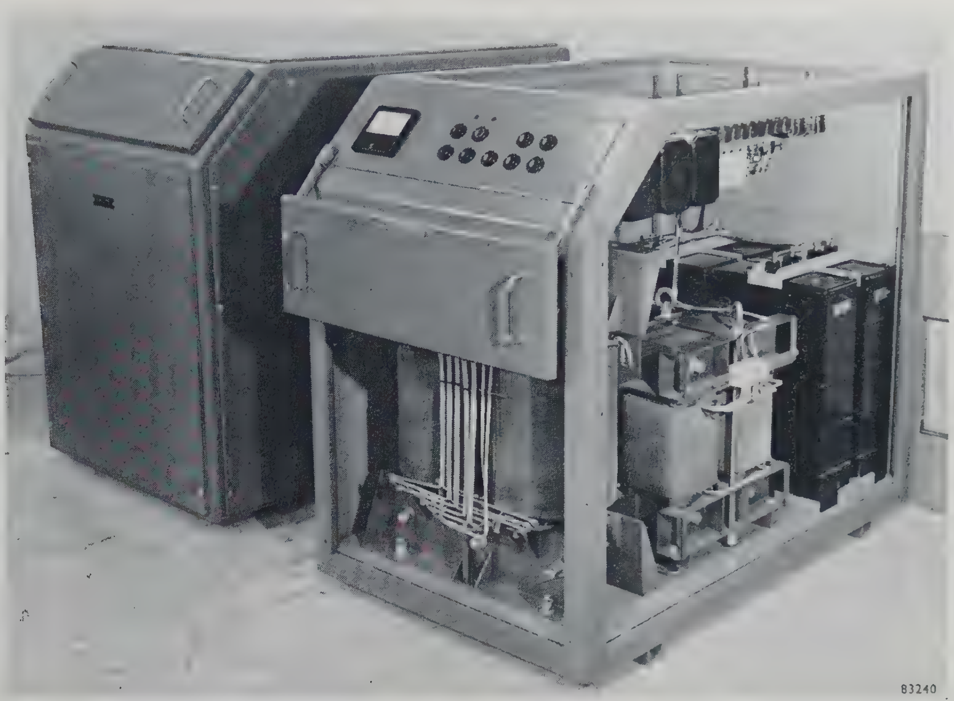

Fig. 10. Two 25 kVA, 20 A supply units, the one on the right with the cover open and the front and sides removed. Note the ammeter and the push buttons on the sloping panel. The isolating transformer, one of the bridge-arm coils and the capacitors can be seen at the bottom, and the magnetic switches at the top of the unit.

cabinet is a control panel (*fig. 10*) with an ammeter to indicate the output current, and push buttons to operate the mains switch, luminous intensity selectors and runway selectors.

The panel is provided with a cover fitted with a door contact to close the remote control circuit; accordingly, local control is possible only with the cover open, and remote control only with the cover closed.

Remote control of the different light systems on an airfield usually operates from the control tower which is then equipped with a panel or desk on

Fig. 11. Control desk. Left: plan of airfield with monitor lamps at points corresponding to the positions of the real runway lights. Right: push buttons and signal lamps of the main switches and the runway and luminous intensity selectors. Centre: panel with clock and meteorological instruments. The spaces on either side are for telecommunications equipment.

which all the light controls are conveniently grouped. Monitor lamps enable the operator to see at a glance which of the systems are in use, and with what luminous intensity, at any given moment. The efficiency of this arrangement can be enhanced by mounting the monitor lamps on a plan of the airfield at points corresponding to the positions of the real lights.

A control desk so arranged is shown in fig. *11*.

The lamp transformers

As stated in the introduction, each lamp in a series system is connected across a separate transformer. The electrical properties of these transformers are highly important if full advantage is to be taken of the special features of the series system, and the reliability of the installation as a whole depends to a very large extent upon their mechanical design.

Electrical characteristics

As regards the electrical properties, firstly, the current in the lamp must remain as far as possible constant, despite a certain amount of spread in the lamp voltage as between individual lamps of the same type and the possibility of contact resistances in series with the lamps. Accordingly, the transformers must possess the characteristic typical of current transformers.

Secondly, the voltage across the lampholder must not build up unduly when a lamp fails or is removed from its holder; otherwise, the replacing of a lamp would involve a certain amount of danger.

Both these requirements are fulfilled by a transformer having the characteristic shown in *fig. 12* (secondary voltage plotted against secondary current). The curve is so steep at the working point that a 10% additional resistance in the secondary circuit reduces the current by only about 0.5%; moreover, the no-load voltage is limited to 3 times the normal operating voltage, this being accomplished by ensuring that the iron transformer-core is completely saturated when there is no load.

However, this saturation is associated with distortion of the primary current; hence the output current of the supply unit is distorted whenever the load is withdrawn from one or more of the lamp transformers. Experience has shown that this increases the r.m.s. value of the current, which is, of course, undesirable, in that, if several lamps happen to fail in a relatively short time, the current in the circuit may increase considerably and so impose a heavy overload on the remaining lamps. Although at civil airports, where the lamps are inspected regularly and defective ones are quickly replaced, this effect is not very important, on military airfields the possibility that many lamps will fail simultaneously, say as a result of enemy action, must certainly be taken into account. Here, then, the authorities will probably require that the r.m.s. current in the series circuit be made as far as possible independent of the percentage of defective lamps. This can be accomplished by detuning the bridge circuit slightly (i.e. by allowing p to deviate slightly from 1), a method which, as we have already seen, is also effective in eliminating the indeterminacy of the arm currents. However, good matching of the supply units to the lamp transformers is then essential, since the proper detuning of the bridge depends upon the transformer characteristic.

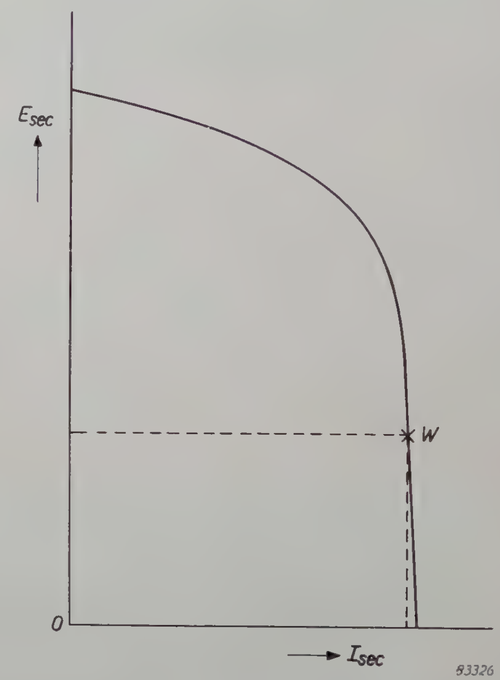

Fig. 12. The secondary voltage (E_{sec}) plotted against the secondary current (I_{sec}) of a lamp transformer loaded with a variable resistor. The primary winding of this transformer is included in a series circuit supplied by a Boucherot bridge. The curve should be steep at the working point W, and the voltage at $I_{sec} = 0$ should not be unduly high.

Mechanical construction

The lamp transformers are designed to operate with lamps ranging from 30 to 500 W, and vary in design according to the particular purpose.

Transformers of the simplest type may be employed for building into runway lights or mounting in a closed box above ground level. The transformer, completely enveloped in moisture-repelling compound, is then placed in a metal box provided with ceramic terminal insulators (*fig. 13*).

Fig. 13. Three transformers of the simplest type mounted in a flush runway light (as shown in fig. 21 of the article referred to in note [1])). The cover, only the bottom of which is seen in the photograph, contains one omni-directional top light and two beam lights (one for each landing direction).

Fig. 14. Taxiway light, mounted on a transformer with cast iron case filled with compound. Note bushings for connection to armoured cables.

Fig. 15. 200 W, 6.6 A lamp transformer for circuits operating with 5000 V supply. It is completely enclosed in rubber, moulded in one piece with the insulation of the leads. The secondary lead (passing through the curved pipe) is fitted with a lamp socket. If the light is knocked over, the safety joint, that is, the small weakened pipe seen in front of the transformer, breaks (see page 282 of the article referred to in note [1])) and the plug is pulled out of the secondary socket. The leads on the extreme right and left are the primary leads. Note the end sleeves employed to protect them from corrosion in storage or in transit.

However, the requirements imposed on transformers to be mounted in pits close to the runway lights, or completely buried, are more stringent. The covering must then resist the action of corrosive elements in the ground; hence the transformer is placed in a heavy, cast-iron box with a removable cover, and the outside of this box is coated with a protective asphalt compound. The transformer itself is delivered from the factory completely enveloped in compound, that is, leaving only the porcelain terminal insulators exposed. When it is installed on the airfield, the cables are connected, the cover is secured, and the remaining space inside the box is filled with compound. When non-armoured cable insulated with a synthetic rubber such as "Neoprene" is employed, the transformer box is provided with coupling nut bushings; for armoured cables, bushings of the type usually employed with cable sleeves are fitted (*fig. 14*).

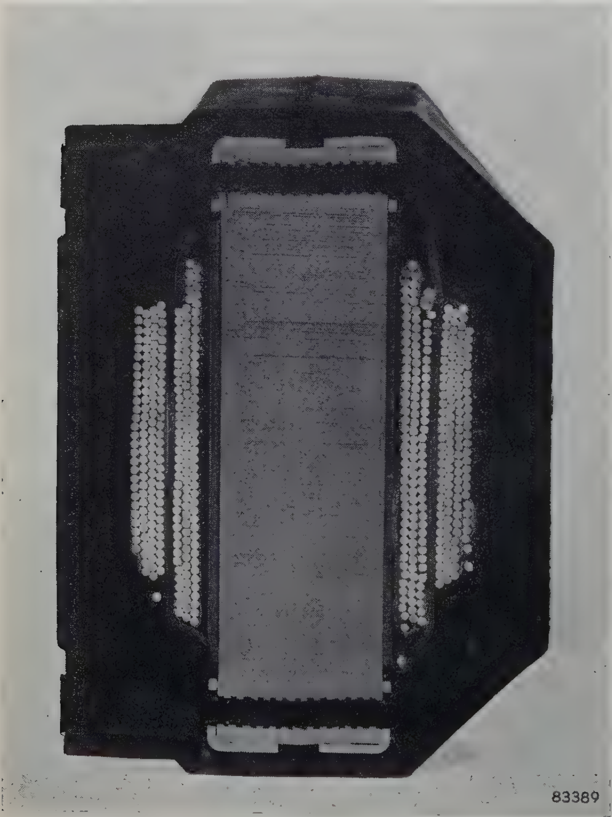

Fig. 16. Cross-section of the rubber-cased transformer shown in fig. 15. It will be seen that the turns are completely enclosed in rubber, which has penetrated into the spaces between them.

Another transformer for use underground, recently developed in America, is hermetically sealed in moulded rubber [5]). Transformers of a similar type are marketed by Philips (*fig. 15*). Here, the waterproof rubber cover is moulded in one piece with the rubber insulation of the leads, which are short lengths of cable with plug sockets vulcanised on to the ends. The transformer is thus protected against corrosion of any kind, and nevertheless convenient to handle. In fact, the mounting of such transformers is so quick and simple an operation that it can be entrusted to unskilled personnel. *Fig. 16* shows a cross-section of the transformer.

The quality of this protective covering is illustrated by the result of a special test to which it was subjected. This test consisted in raising the transformer daily to the working temperature and then cooling it abruptly by plunging it in water. Although this very stringent test was continued for several weeks, insulation resistances of the order of thousands of megohms were afterwards measured at a test voltage 3 times the normal operating voltage of the transformer.

[5]) L. C. Vipond, No drinks for transformers, Aviation Age **17**, 56, November 1952.

Summary. For airfield light beacons incandescent lamps of voltages ranging from 6 to 30 V, each connected to a separate transformer, are employed. In the present article it is shown that such transformers are better connected in series than in parallel, since in a series circuit the voltage drop in the supply cables causes no variation in luminous intensity between the individual lamps connected to the transformers; hence it is possible to employ very much thinner cables and so considerably cut down on cable costs, which constitute an important part of the total capital investment. Another advantage of the series system is that it is more reliable (effect of contact resistances less noticeable; separate lamp fuses may be dispensed with).

The Boucherot bridge circuit is a suitable current source for a series system. It has two convenient features; the output current does not depend upon the nature or size of the load impedance, and with resistive loading the power factor is unity.

However, the indeterminacy of the individual currents in the four bridge arms is a disadvantage. Measures to eliminate this include slight detuning of the bridge.

Philips employ the Boucherot bridge in a series of airfield light-beacon supply units ranging from 1.5 to 25 kVA. Certain features of these units, including the method of limiting the output voltage when the external circuit is interrupted, and a new interlock system, are described.

Finally, the lamp transformers are discussed. The characteristic of such a transformer must precisely match that of the associated supply unit. One type of transformer, insulated entirely with rubber, is specially good for its corrosion resistance.

ELECTRONIC CONTROL OF INDUSTRIAL PROCESSES

by H. J. ROOSDORP.

621-523.8:621.317.733.083.4.078

Referring to a previous article giving a general review of regulating systems for industrial processes, we shall deal here with the various forms of electronic controllers and their advantages with respect to non-electronic control systems.

Introduction

In many industrial processes the quality of the final products is dependent on various physical or chemical quantities, e.g. temperature, pressure, acidity, etc. In such processes it is necessary that these quantities are adjusted to the most favourable value and that this adjustment is maintained by checking and controlling the production process in a suitable manner. This can be done either manually by an operator or automatically, without human intervention. In the former case the quantity to be regulated has to be measured and read from some scale before it can be adjusted to its correct value. Strictly speaking such measurements — or at least such readings — are not necessary when the process is automatically controlled. If the installation is functioning properly, the value of the controlled quantity should remain practically equal to what is desired. In fact, there are a great number of such non-indicating controllers, which operate without showing the value of the controlled quantity. In most cases, however, indication of the controlled quantity is desirable, since it gives a continuous check on the progress of the operation. Indeed it is often convenient to have a record of the controlled quantity as a function of the time; recording controllers are therefore frequently used.

An electrical instrument (automatic potentiometer) of robust construction, for measurement and recording in industrial processes has been described earlier in this Review [1]). The present article deals with some of the ways in which this instrument can be used for automatic control. For a study of the general aspects of automatic control we may refer back to another article in this Review [2]). For the sake of simplicity, the present article is mainly concerned with one specific process, viz. maintaining the temperature of a gas-fired furnace. The discussion, however, is equally valid for entirely different processes.

In the automatic control of furnace temperature a "closed loop" is formed by the furnace, the temperature-measuring instrument, the controlling unit and the correcting element (in this case the valve regulating the gas supply). In most cases an actuating mechanism of some sort will be required between the automatic controller and the valve, e.g. an electric motor or an electro-pneumatic valve positioner. *Fig. 1* gives a schematic representation of the closed loop thus obtained. The controlling unit Q establishes the functional relationship between the temperature of the measuring element Th and the position of the gas valve K, i.e. the "control action" of the system. A controller of this kind may be mechanically, electrically, hydraulically or pneumatically operated. Each of these systems has its specific merits and drawbacks. Electric and more particularly, electronic systems are finding increasing application. This is essentially due to the following reasons. Electricity is almost universally available, so that no other sources of power have to be installed (e.g. compressed air supply). Furthermore, most measurements are in any case carried out electrically, so that it is often simple to extend the measuring circuit with a

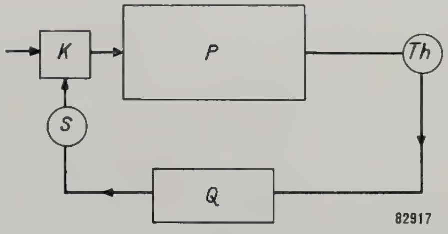

Fig. 1. Diagram of a closed loop automatic control system. The process to be controlled is represented by P. Th measuring element; Q controller; S motor element; K correcting element.

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1) H. J. Roosdorp, An automatic recording potentiometer for industrial use, Philips tech. Rev. **15**, 189-198, 1953/54.

2) H. J. Roosdorp, On the regulation of industrial process, Philips tech. Rev. **12**, 221-227, 1950/51.
See also, e.g., W. Oppelt, Kleines Handbuch technischer Regelvoränge, Verlag Chemie, Weinheim, 1954; and D.P. Eckman, Principles of Industrial Process control, John Wiley, New York 1945.

circuit for electrical control. Transmission of electric signals is virtually instantaneous, which may be important if large distances (e.g. in excess of 100 m) have to be bridged. The elements used in electric circuits (mainly resistors and capacitors) to obtain the required control action have substantially constant values, so that the functioning of an electric controlling circuit is restricted to linear operations involving simple mathematics. Moreover, these circuit elements are standard radio components, everywhere obtainable and easy to handle. The design of the amplifiers used in these circuits may draw on the experience gained in telecommunications. For these and other reasons, the control action with an electronic controller is achieved in a relatively simple manner. As a final argument it should be mentioned that the maintenance and testing of electric and electronic instruments and, if necessary, the replacement of parts, are usually simple routine work.

Sometimes it is desirable that the position of the valve depends not only on the deviation of the measured quantity from the desired value, but also on the time for which this deviation has existed, or on the rate at which the deviation increases or decreases (integral or derivative control respectively, cf. [2])). Futhermore it may be necessary that the desired value of the controlled condition changes as a function of time (programme control). All these requirements can be satisfied by electronic controllers in a simple way.

If no work of a mechanical nature is required of the correcting element, a wholly electric system can be used. An example of this is the regulation of an electric furnace with the aid of transductors or gas-filled tubes. If, however, as in the case considered here, the position of a mechanical valve has to be altered, a motor element (valve motor S in fig. 1) has to be used, which may operate electro-mechanically, pneumatically or hydraulically.

In the following we shall not be concerned with the nature of this mechanism, but only with the electrical part of the controlling unit.

Automatic control by means of the automatic potentiometer

The working principle of the automatic potentiometer/bridge is as follows: a contact fitted on a carriage is moved by a motor along a resistance slidewire incorporated in the measuring circuit, the motor adjusting this contact to a position corresponding to the value of the measured quantity. For automatic control, a voltage is required whose amplitude and polarity corresponds to the deviation of the measured quantity (in this case the furnace temperature) from the desired value. Such a voltage can be obtained by means of a second slidewire, equal in length to the measuring slidewire and with a movable contact fixed to the same carriage as that on the latter. On the controlling slidewire, a second contact is fitted which can be set by hand to a position corresponding to the desired value of the controlled condition. With a voltage E_1 applied across the controlling slidewire, there is a voltage E_x between its two contacts (*1* and *2* in *fig. 2*) that is proportional to the difference between the actual furnace temperature and the desired furnace temperature. This voltage can be used to determine, via an actuating mechanism, the position of the gas valve.

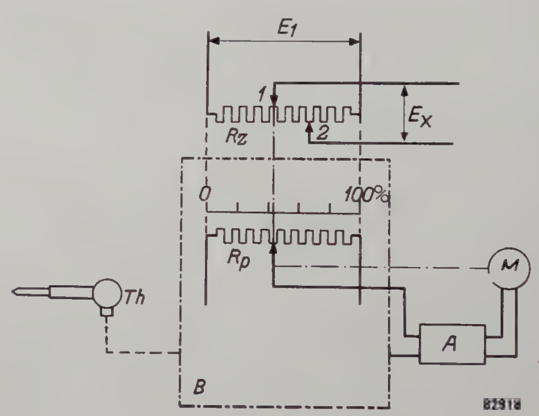

Fig. 2. Use of the automatic potentiometer for control of the temperature of a furnace. B measuring circuit; Th thermo-element; R_p measuring slidewire; A amplifier; M drive motor; R_z control slidewire. The contact *1* is mounted on the same carriage as the contact of the measuring slidewire. Contact *2* is set to a position corresponding to the desired temperature.

In the following we shall deal with some circuits for electronic controllers and what can be achieved with them in industrial processes.

Two-step control

One of the simplest systems of control is that in which the valve or correcting element has two positions only, viz. a position A, which, if maintained, would raise the furnace temperature T to above the desired value T_0, and position B which would cause the temperature to drop below T_0. The valve has to be in position B if T is higher than T_0 and in position A if T is lower than T_0. Each time the desired value is passed, the valve changes its position, so that the temperature continues to oscillate between two values, one higher and one lower than T_0.

This can be effected with the circuit shown in *fig. 3*. The position of the contact *2* of the controlling slidewire R_z corresponds to the desired value of the temperature; that of contact *1* to the actual

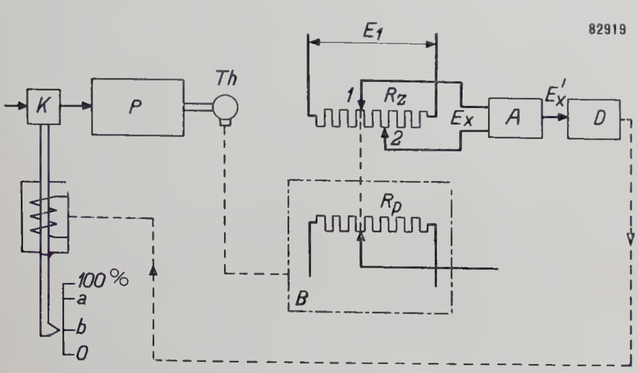

Fig. 3. Circuit for two-step control of the temperature of a furnace. *A* amplifier; *D* phase-sensitive detector. The other letters have the same significance as in figs. 1 and 2.

value of the temperature. An alternating voltage E_1 of mains frequency, is applied to the controlling slidewire. The voltage E_x between the contacts *1* and *2* (the error voltage), after being raised to a value E_x' by the amplifier *A*, is applied to a phase sensitive detector *D*. The circuit diagram of the latter is shown in *fig. 4*. The alternating voltage E_x' is added to an alternating voltage E_k and applied to the grid of ~~a~~ triode. An alternating voltage E_a is applied to the anode of this tube. E_a, E_k and E_x' have, of course, the same frequency. E_k and E_a are in anti-phase and have such values that, when E_x becomes zero, anode current ceases to flow through the triode. E_x' and E_a are either in phase or in anti-phase, according to whether the contact *1* is to the left or to the right of contact *2*. If both voltages are in phase, anode current flows through the tube during each half-period and the armature of relay *Re* is attracted.

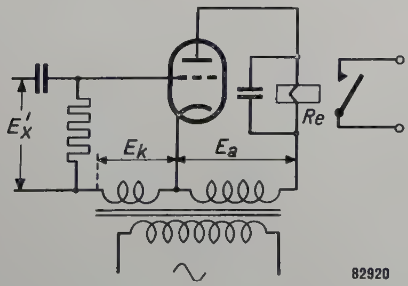

Fig. 4. Diagram of a phase sensitive detector. *Re* Relay.

If on the other hand E_x' is in anti-phase with E_a, the tube passes no anode current and the relay is not energised. Whether or not the relay is energised is thus an indication of the direction in which the temperature deviates from the desired value. This relay operates the correcting element, e.g. a gas valve, via a motor element and sets it to either position *A* or position *B*. The circuit is preferably so arranged that the releasing of the armature of *Re* corresponds to a safe position in the process to be regulated; in the present example, therefore, it corresponds to the closed position of the gas valve. Should the controller break down, so that E_x' remains zero, there is then no chance of the temperature rising to a dangerous value.

The positions *A* and *B* of the gas valve correspond to *a* % and *b* % respectively ($a > b$) of the maximum opening y_{max} of the valve. The inherent temperature oscillations are smaller according as the difference between *a* and *b* is less. In practice, the extent to which this difference can be reduced is restricted by the fact that at $y = y_{max}\, a/100$ the temperature must be able to rise above T_0 under all circumstances, that is to say, even under the most adverse conditions, e.g. the lowest ambient temperature and the lowest gas pressure. Conversely, at $y = y_{max}\, a/100$ the furnace temperature must be able to drop below T_0 even at the highest ambient temperature and the highest gas pressure. The neccessary difference between *a* and *b* is therefore, as a rule, dictated by local conditions. In a special case *a* can be 100 % and *b* 0 %. Such a system is termed on-off control. A system like this obviously involves fairly large oscillations in the temperature, but it has the advantage that in many cases the construction is relatively simple.

The magnitude of the oscillations may be limited by selecting the most suitable position of the measuring element with regard to the source and dissipation of the heat. If the measuring element reacts (via the furnace temperature) to the position of the gas valve with only a slight time lag, excessive fluctuations may be avoided.

Multi-step control

If for a given process the external conditions (ambient temperature, gas pressure etc.) vary to such an extent that a two-step control would necessitate an excessively large differential between *a* and *b*, the following system can be used. The differential between *a* and *b* is made just large enough to ensure a properly functioning control for small fluctuations of the external conditions. A third setting *C* of the valve is provided to take care of those situations in which the small differential between *a* and *b* cannot cope with the wide variation in external conditions. If, for example,

the ambient temperature, were so high that, even with the valve in the B position, the temperature continued to increase, the third position C of the valve would be set at a more closed position than B, i.e. $c<b$. Thus, if the temperature should exceed a certain value T_1 ($>T_0$) the controller causes the valve to move to the c position, where it stays until the temperature has dropped below T_1. The setting c is of course fixed at such a value that under all condition, the temperature T can drop below T_1 when $y=y_{\max}\,a/100$. The net result is that at a high ambient temperature, control will take place about T_1, just as with two-step control. If the ambient temperature falls, control again takes place about T_0. This is known as three-step control. *Fig. 5* is a sketch showing how such a system can

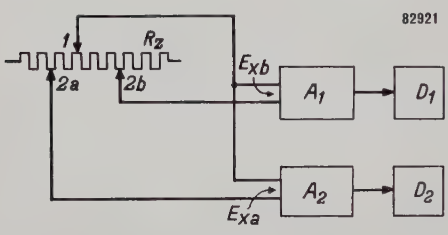

Fig. 5. Part of a circuit for three-step control. R_z controlling slidewire; A_1 and A_2 amplifiers; D_1 and D_2 phase sensitive detectors. Contact *1* moves with the contact of the measuring slidewire. Contacts *2a* and *2b* are adjusted by hand.

be realized. The controlling slidewire R_z is now provided with two manually adjustable contacts, *2a* and *2b*. The contact *1* is again moved together with the contact of the measuring slidewire. The two voltages E_{xa} and E_{xb} are applied, via amplifiers, to separate phase detectors D_1 and D_2. The position of contact *1* determines whether both the phase detector relays are open, both closed or one open and one closed. These three possibilities determine, via a motor element, the three positions of the gas valve. *Fig. 6* shows the effect of a three-step control on the furnace temperature T as a function of the time t when the ambient temperature is rising. Initially T oscillates regularly about the value T_0, but after some time it rises and oscillates about T_1. The three-step control discussed is the simplest case of a multi-step control. Clearly this system may be extended in principle to a control system with any number of steps. With multi-step controls, oscillations in the furnace temperature due to large changes in external conditions can be kept reasonably small.

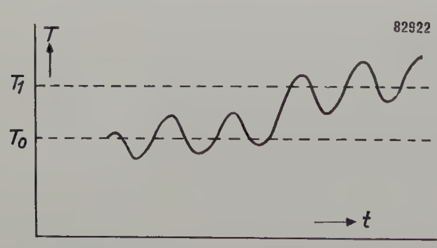

Fig. 6. The temperature T of a furnace as a function of the time t for a system of three-step control, the external conditions being such that the furnace temperature has a tendency to rise continuously.

Proportional control

Both two-step control and multi-step control are examples of discontinuous control systems. In these systems the gas valve (or any other correcting element) can be set only to a number of discrete positions. This limitation does not exist in continuous control systems. One such continuous control is proportional control. This method may be considered as an extreme case of a multi-step control system with an infinite number of steps at infinitely small intervals. The displacement of the correcting element is then proportional to the deviation of the furnace temperature from the desired value. (The time required for the movement of the correcting element is disregarded here.)

Proportional control could be realized electrically with a gas valve operated by a solenoid to which a voltage is applied proportional to the difference between the actual and the desired temperature. This construction, however, has the drawback that the relationship between the valve opening and the voltage applied to the solenoid is likely to be disturbed by forces acting upon the valve (e.g. frictional forces). A further drawback is that considerable power must be expended merely to maintain the position of the valve.

A system less subject to these drawbacks is shown schematically in *fig. 7*. Here again we have the controlling slidewire, R_z, supplied with a voltage E_1. The contact *1* of R_z is moved by the carriage of the measuring slidewire (not shown in the diagram). The voltage E_x tapped off on this slidewire is added to a second voltage E_y, which is obtained from the two parallel slidewires R_y and R_c, supplied with a voltage E_2. The contact *4* of R_y is mechanically linked to the correcting element (gas valve). Contact *3* of R_c can be set by hand. E_1 and E_2 are alternating voltages derived from the mains via a transformer.

The difference between the two voltages E_x and E_y is applied to the amplifier A_1. The output voltage of the latter controls the motor M_1 of the valve in such a way, that owing to the displacement of the

contact *4*, the voltage difference $E_x - E_y$ becomes zero. Since (see fig. 7)

$$E_x = (x - x_0)\, E_1$$

and

$$E_y = (y - y_c)\, E_2 ,$$

it follows that

$$y - y_c = \frac{E_1}{E_2}\,(x - x_0) \quad . \; . \; . \; . \; . \quad (1)$$

The displacement of the gas valve with respect to the position corresponding to $y = y_c$ is thus proportional to the distance between contact *1* and *2*. Since there is usually a linear relationship between the furnace temperature and the displacement of contact *1*, proportionality between the displacement of the valve and the temperature deviation of the furnace is thus achieved. The proportional control factor E_1/E_2 can be varied by the choice of E_1 and E_2.

In general, x can reach the value x_0 only when y_c is made equal to the valve position y. This condition, however, is possible only if the external conditions are such that for $y = y_c$ the furnace temperature does indeed become equal to the desired value. This can be realized by adjusting y_c by means of contact *3*. If, after this, any change occurs in the process to be controlled or in the external conditions, another position of contact *3* would be necessary in order to make $x = x_0$ when $y = y_c$. If, however, the contact is left at its original setting, then x will differ from x_0 by an amount which remains small so long as the proportional control factor has a large value.

In practice it is not possible to make the proportional control factor arbitrarily large. Because of the inevitable time lag that always exists between a displacement of the gas valve and its complete effect on the furnace temperature, too great a value of the proportional factor may create an unstable condition in which the furnace temperature is subject to large oscillations [3]). This restricts the choice of the proportional control factor. The large drift from the desired value which can occur if the factor is too small can, of course, be corrected by adjusting contact 3, but this may be most inconvenient if it has to be done too frequently. By adding to the proportional control a circuit for integral control, adjustment of contact *3* is no longer necessary, as will be demonstrated presently.

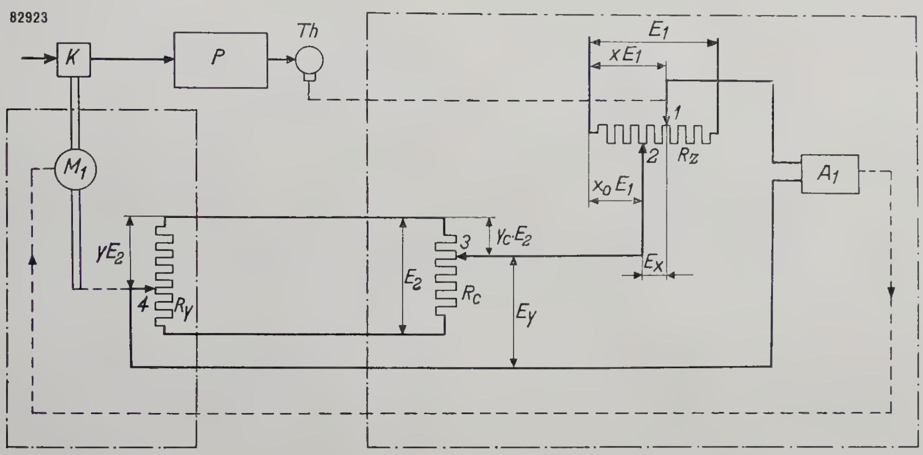

Fig. 7. Schematic lay-out of a proportional control system. The right-hand part (surrounded by chain line) is situated near the measuring element, the left-hand part near the correcting element (valve).

Floating control and integral control

With proportional control there is a fixed relationship between the deviation of the furnace temperature and the position of the correcting element. In the case of floating control, however, the final control element is continuously displaced until the furnace temperature no longer deviates from the desired value. The rate of displacement may be constant, or it may have two or more fixed values, or it may be proportional to the value of the deviation. In the latter case the total displacement of the correcting element is proportional to the time integral of the temperature deviation. This

[3]) This is completely analogous to the fact that an amplifier with negative feedback can go into oscillation if the feedback is large and if a large phase shift occurs at certain frequencies.

special case of floating control is called integral control.

An example of a floating control system is shown schematically in *fig. 8*. The error voltage E_x is

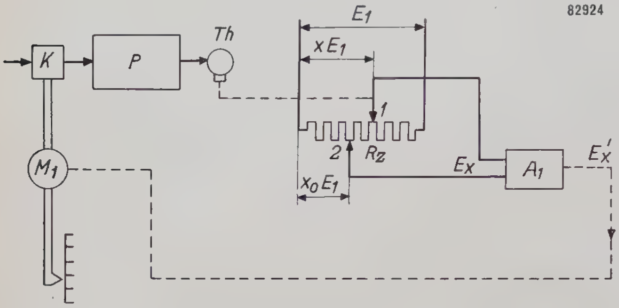

Fig. 8. Scheme for floating control. The motor M_1 actuated by the output voltage E_x' of the amplifier A_1, changes the position of the correcting element K.

applied to the amplifier A_1. The output voltage E_x' of this amplifier controls the motor M_1 by which the final control element is driven. If the speed of the motor is made proportional to the applied voltage, an integral control is obtained (also known as proportional speed floating control).

The circuit may also be so arranged that the speed of the motor remains constant (single-speed floating control). In this case only the direction of rotation of the motor, is governed by the direction of the furnace temperature (i.e. on the sign of of the deviation $x - x_0$). This can be effected by applying the output voltage of the amplifier A_1 to a pair of phase detectors, whose supply voltages are in anti-phase. (*fig. 9*). Whether relay Re_1 or relay Re_2 is energized depends on the sign of $x - x_0$, so that the motor is switched on with the desired direction of rotation. If $x - x_0 = 0$, which means that the controlled variable has the desired value, both relays are without current and the motor remains at rest.

With floating control the correcting element is at rest only when the deviation of the furnace temperature is zero. It therefore avoids the disadvantage of proportional control, namely, that in spite of the control action, there remains a temperature deviation. An advantage of proportional control, however, is that once a change in the controlled quantity occurs, the valve or correcting element very quickly moves to the position corresponding to the new situation. (The time of adjustment is usually limited only by the speed of the valve itself.) In the case of floating control, the speed of the motor element is relatively slow. This means that if the changes in the controlled quantity occur rapidly, the final control element cannot quickly enough reach the position necessary to establish zero deviation of the controlled variable from the desired value. Thus also with floating control, a sudden change in the process results temporarily in excessive deviations of the controlled quantity. Raising the speed of the motor element may throw the system into oscillation. For this reason floating control is generally used in combination with proportional control. A floating control of the integral type is usually chosen in such cases.

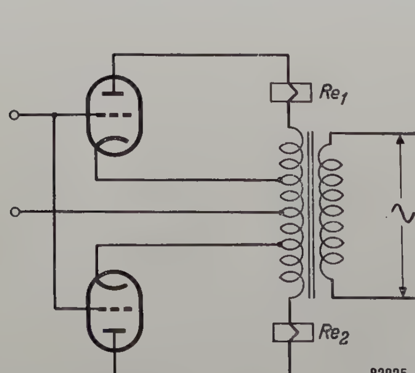

Fig. 9. Combination of two phase detectors, for use with a floating control with constant motor speed.

Proportional and integral control

Fig. 10 shows a simplified diagram of a control system in which the displacement of the final control element (y) depends both on the deviation from the furnace temperature (therefore upon $x - x_0$) and on the time-integral of this deviation. The voltage E_y, which depends on the position of the correcting element, is now not connected directly in series with E_x, but via the capacitor-resistor coupling C-R. The voltage E_1 and E_2 are direct voltages in this circuit [4]). Again, the correcting element is adjusted in such a way that the input voltage to the amplifier becomes zero. From fig. 10 we have:

$$E_x = iR.$$

Furthermore

$$E_y = E_2(y - y_c) = \frac{1}{C}\int i\,\mathrm{d}t + iR$$

and

$$E_x = (x - x_0)\,E_1.$$

From these equations it follows:

$$y - y_c = \frac{E_1}{E_2}\left\{(x - x_0) + \frac{1}{RC}\int_0^t (x - x_0)\,\mathrm{d}t\right\}. \quad (2)$$

[4]) The direct voltage applied in this case to the amplifier A_1 will usually first be converted by means of a vibrator-converter into an alternating voltage, so that a normal A.C. amplifier can be used (Cf. e.g. Philips tech. Rev. **16**, 117-122, 1954/55, (No. 4).

In this way we see that the displacement of the correcting element can be separated into two terms, viz. a term $(x-x_0)E_1/E_2$, which is proportional to the deviation from the furnace temperature, and a term that is proportional to the time-integral of this deviation. The latter term will be larger the smaller the value of the product RC (it is assumed that the furnace reacts so slowly that during the interval considered $x-x_0$ remains virtually constant or that contact 1 is kept temporarily in a fixed position). Adding the two terms we see that the initial change in $y-y_c$ due to the proportional control action is doubled by the integral term after a period of time given by

$$t = RC.$$

This is called the integral action time.

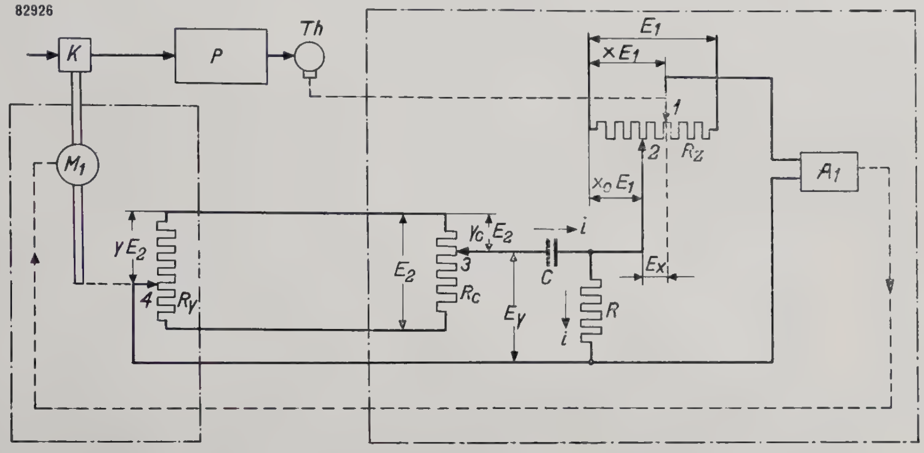

Fig. 10. Circuit for proportional + integral control. By interchanging the elements C and R, a circuit for proportional + derivative control is obtained.

The significance of this RC-product may be clarified by the following considerations (*fig. 11*).

If, starting from the condition where $x = x_0$, a sudden change in x occurs, this is accompanied by a sudden change in y to the effect that $y-y_c = (x-x_0)E_1/E_2$. Because of the integral term in (2) however, $y-y_c$ is further increased by a term

$$\frac{E_1}{E_2} \cdot \frac{x-x_0}{RC}\, t$$

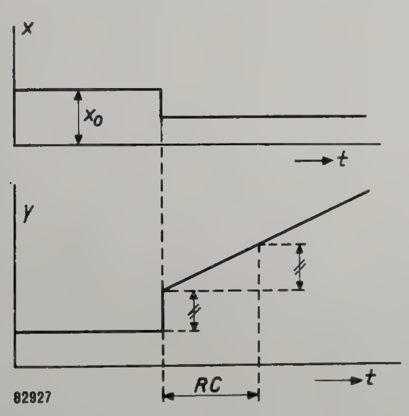

Fig. 11. Variation of x and y with time, due to a sudden change in the controlled variable x, for a system of combined proportional and integral control.

From (2) it is self-evident that the correcting element is displaced as long as there is any deviation present between x and x_0. Hence, with the system considered here, the controlled variable does in fact attain the desired value.

Proportional and derivative control

In our discussion of proportional control we have already pointed out that the proportional control factor cannot be made indefinitely large because of the risk of instability. In many cases the stability may be improved by using a circuit in which the correcting element is given a displacement proportional to the time derivative of the deviation from the controlled condition in addition to the displacement due to proportional action. A circuit suitable for this can be derived from fig. 10 by interchanging the capacitor C and the resistor R. By a similar argument to that used for integral control it can be shown that the displacement of the final control element is given by the equation:

$$y-y_c = \frac{E_1}{E_2}\left\{(x-x_0) + RC\frac{\mathrm{d}}{\mathrm{d}t}(x-x_0)\right\}. \quad . \quad (3)$$

Here too, the latter term depends upon the RC-product. The significance of this product can be established as follows. Suppose that the system, initially in equilibrium ($x = x_0$), is subjected to a change such that x varies at a constant speed,

i.e. dx/dt is constant. Let the value of dx/dt be k (see *fig. 12*). The correcting element is then given a sudden displacement, due to the derivative action, of value $y - y_c = RCkE_1/E_2$. It also gets a displacement from the proportional action, and this increases with time since the variable itself $(x - x_0)$ has been assumed to be changing linearly with time; this displacement is given by $y - y_c = ktE_1/E_2$. These two displacements reach the same value after a time $t = RC$. This period is called the derivative action time.

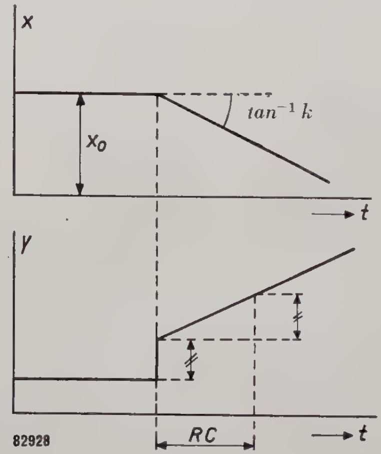

Fig. 12. Variation of x and y due to a change in x at constant rate for a system of combined proportional and derivative control.

Proportional, integral and derivative control

The properties of the two previous circuits can be combined as in the circuit of *fig. 13*. The two RC-networks, R_1C_1 and R_2C_2, are now both incorporated. The displacement of the correcting element is then given by the equation

$$y - y_c = \frac{E_1}{E_2}\left\{(x - x_0)\left(1 + \frac{R_2}{R_1} + \frac{R_2C_2}{R_1C_1}\right) + \frac{1}{R_1C_1}\int_0^t (x - x_0)\,dt + R_2C_2\frac{d}{dt}(x - x_0)\right\}. \quad (4)$$

The integral and derivative actions are given by the same terms as in equations (2) and (3), but the proportional action is now given by the term

$$\frac{E_1}{E_2}(x - x_0)\left(1 + \frac{R_2}{R_1} + \frac{R_2C_2}{R_1C_1}\right),$$

so that in this case, the proportional action also depends on the value of R_1, C_1, R_2 and C_2.

Time-proportional control

The continuous control of large electric furnaces requires large and elaborate control devices, such as thyratrons, variable transformers and transductors. For this reason a discontinuous control action is often preferred, in which the furnace is simply switched on and off at intervals. A far simpler apparatus (as a rule a magnetic switch) can than be used, but this has the disadvantage that quite considerable oscillation in the furnace temperature may occur. It is, however, also possible to obtain the effect of a continuous control, whilst maintaining the simplicity of a discontinuous control. This can be achieved by switching the applied power either on and off or between two specific values, at a constant frequency. The ratio between the periods during which the furnace is switched on and switched off is made continuously variable and dependent upon the difference between the actual and the desired furnace temperature. The relationship between these

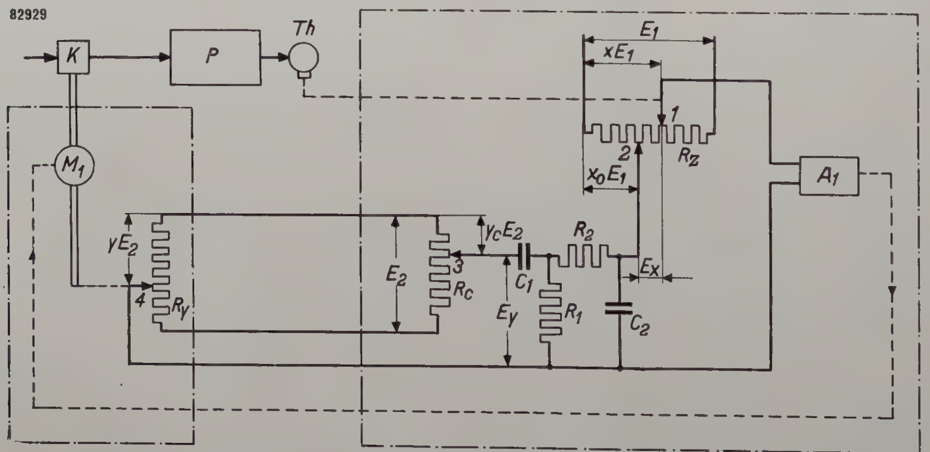

Fig. 13. Layout of circuit for proportional, integral and derivative control.

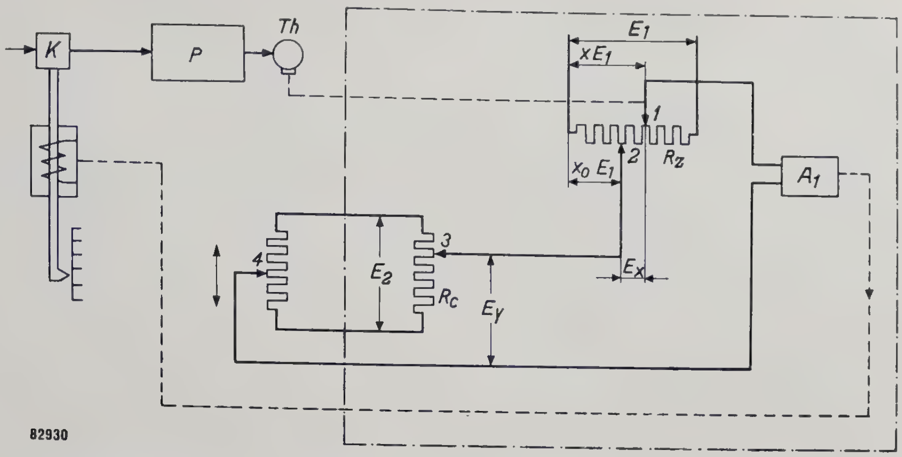

Fig. 14. Diagram of a circuit for control by time-proportional switching. The contact *4* is moved up and down periodically at a constant rate.

two values may be proportional, integral or derivatative, or it may be a combination of these.

Fig. 14 shows a simplified diagram of a circuit for obtaining a proportional control in this way. Here again the voltage E_y is added to the voltage E_x of the controlling slidewire. Contact *4* is now moved periodically up and down, so that the voltage E_y shows a similar periodic variation. In *fig. 15* the voltages E_x and E_y are plotted as func-

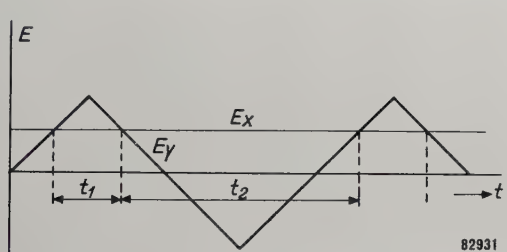

Fig. 15. Voltages E_x and E_y of fig. 14 as functions of the time. t_1 and t_2 represent the times during which the furnace is switched on and switched off, respectively.

tions of time. As regards the latter voltage, if contact *3* is placed in the middle, E_y will vary symmetrically, about the zero line. The instants when the input voltage of the amplifier passes through zero are given by the points of intersection of the E_x and E_y curves. At these instants the input voltage changes its phase by 180° which, by means of a phase detector and a magnetic switch, causes the furnace to be switched alternately on and off. As can be seen from fig. 15, the ratio of the time t_1 during which the furnace is switched on, to the time t_2 during which it is switched off, is dependent upon E_x. The mean value of the applied power is thus continuously dependent upon the deviation of the furnace temperature. The rate at which switching occurs (i.e. the period of the E_y oscillation) should be so selected that the fluctuations of the furnace temperature caused by it, are sufficiently small.

As in the system of proportional control dealt with above, a certain deviation from the controlled condition after a change in the process will likewise remain with a circuit such as that shown in fig. 14. Here too, this can be corrected if necessary, by re-adjusting contact *3*.

Programme Control

In the foregoing the value of the controlled condition was assumed to be constant. In some cases, however, it is necessary to vary the desired value as a function of the time. One way in which this could be achieved is by incorporating in the control unit a mechanism by which contact *2* is displaced in accordance with a given time function, viz. the "programme" of the process. Generally, however, a method is preferred which does not require any special mechanism to be built into the controlling unit. An example of this is the circuit shown in *fig. 16.* A separate apparatus, the programme transmitter, contains a second slidewire R_p in parallel

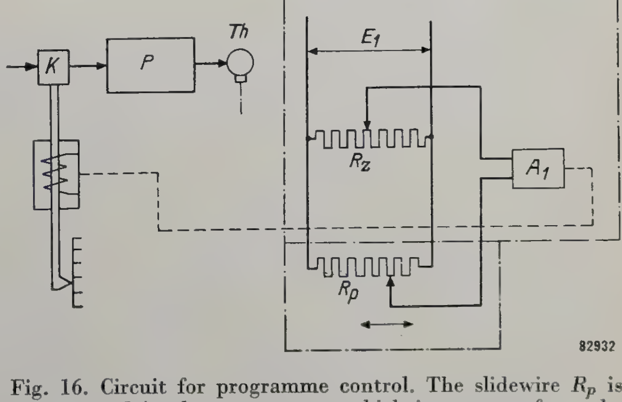

Fig. 16. Circuit for programme control. The slidewire R_p is incorporated in the programmer which is separate from the potentiometer.

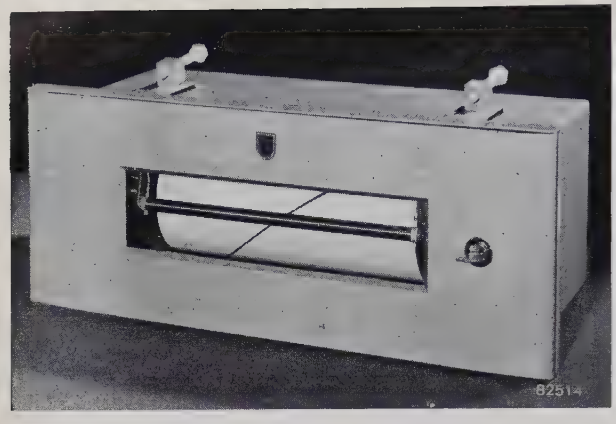

Fig. 17. The Philips programmer, type PR 7211.

with the controlling slidewire R_z, The contact *2*, which serves for setting the desired value is now mounted on this second slidewire R_p and is moved along it in accordance with the required time function. The Philips programme controller has a rotatable cylinder of insulating material mounted parallel to R_p, carrying a conducting wire on its surface running along the graph of the required time-function. At one spot this wire is in contact with the potentiometer R_p. The cylinder is rotated at uniform speed by an auxiliary motor, so that the point of contact moves according to the desired time function. The programmer is illustrated in *fig. 17*.

Alternative control systems

A number of the control systems described in the foregoing may also be realized in other ways, still using the automatic potentiometer for the measurement. An on-off control, for example, can be realized by using the carriage of the measuring slidewire to operate an on/off contact at a certain point. There is the drawback, however, that in this case a greater force is necessary for displacing the measuring carriage at the position corresponding to the value of the controlled condition. In this way the sensitivity of the measuring bridge is impaired. Furthermore, for making and breaking a contact, a certain displacement of the carriage is always required, so that there would be a permanent differential between the points of switching on and switching off. In the control system already described, the carriage has to overcome only relatively small frictional forces and the sensitivity of the measuring bridge is uniform. By using a high gain amplifier, the differential between switching on and switching off may be reduced virtually to zero.

Systems for continuous control may also be derived mechanically, pneumatically or hydraulically from the automatic potentiometer. An integrating and derivative system can be realized, for example, pneumatically or hydraulically by means of capillaries and volume elements. These, however, require more maintenance than the corresponding electrical elements (resistors and capacitors). In pneumatic systems, for example, it is necessary to use dry, oil-free and dust-free air to minimize rust and clogging or freezing in orifices and capilliaries.

The use of an electric control system does not exclude the use of hydraulic or pneumatic final control elements. An electrically controlled hydraulic or pneumatic valve positioner or similar device is then used.

Summary. A survey of the various types of automatic control actions and how these may be achieved electronically for the control of industrial processes. The various circuits described are suitable for use in conjunction with the automatic potentiometer/bridge described in a previous article. The various methods of control may be divided into systems for discontinuous and for continuous control. The discontinuous system most widely used is the two-step control. If the conditions under which the process takes place vary considerably, this method can be extended to a three-step or multi-step control. When the number of possible positions of the regulating unit is made very large, this system becomes a continuous control system, e.g. proportional control. One objection to the latter is that the controlled variable may not remain exactly at the desired value. This objection may be overcome by using a floating control system, of which integral control is a special form. The drawback inherent in integral control, viz. that the correcting element does not react instantaneously to a sudden disturbance in the process conditions, may partly be obviated by combining integral and proportional control. In order to improve the stability when the proportional control factor is large, derivative action may be added to proportional control. The quasi-continuous control of electric furnaces may be achieved with advantage by means of time-proportional switching; with simple circuitry the ratio of the on and off times of the furnace may, for example, be made proportional to the deviation of the furnace temperature from the desired value. Finally, a brief description is given of programme control and of alternative non-electronic control systems used in conjunction with an electronic automatic potentiometer.

LINEAR ELECTRON ACCELERATORS FOR DEEP X-RAY THERAPY

621.384.62:615.849

Along with the development of linear electron accelerators for nuclear physics research, similar machines have been designed and built for medical purposes. Linear accelerators are capable of producing an intense narrow beam of electrons having energies of several millions of electronvolts [1][2]. When impinging on a suitable target, the electrons generate X-rays of high penetrating power, so providing a very efficient source for deep therapy.

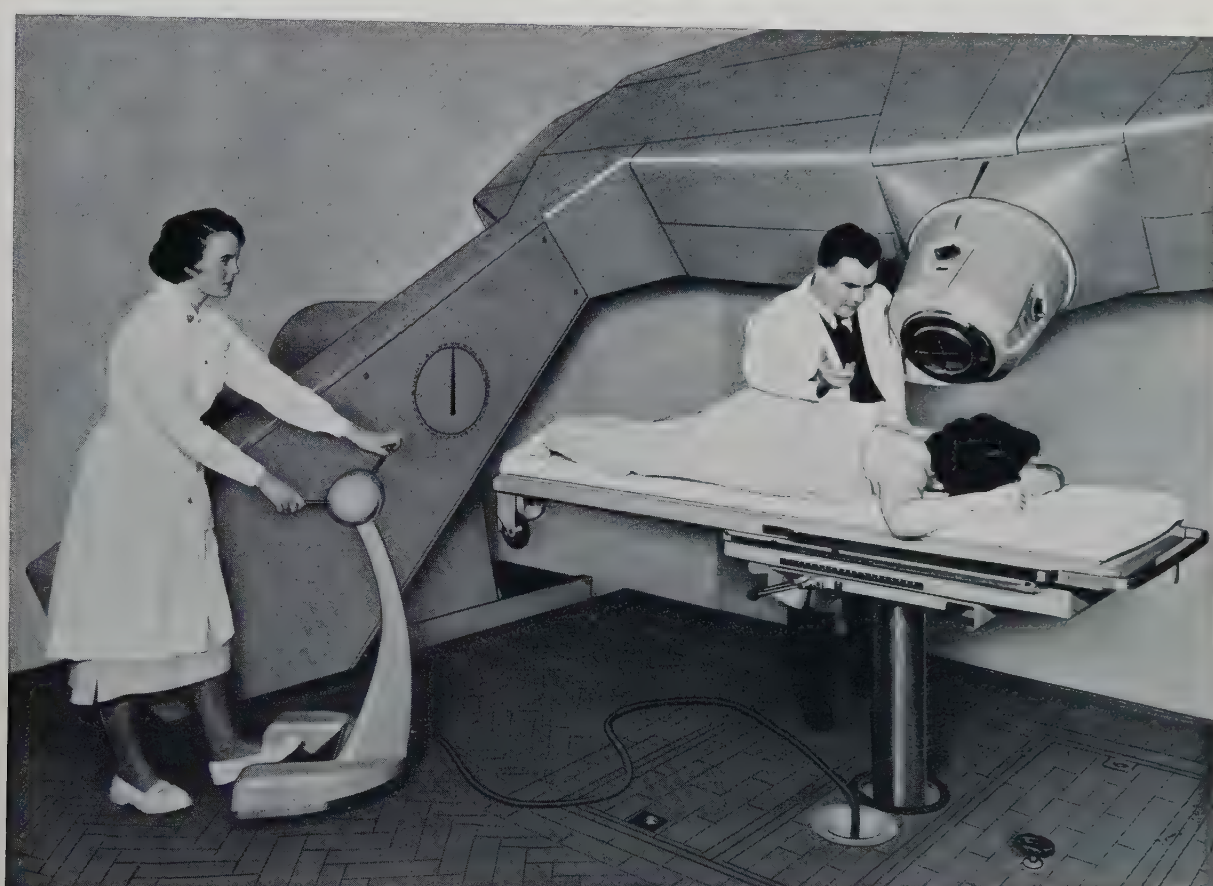

Fig. 1. Medical linear accelerator generating 4 MeV X-rays, installed at Newcastle Genera Hospital (England). This was the first of a series of 4 MeV-accelerators to be used for deep X-ray therapy.

The photograph shows a patient being set up for treatment. The double-ended gantry houses the accelerator with accessories and X-ray head. The assistant on the left adjusts the height of the couch and the angle of the gantry to provide the correct point and angle of incidence of the radiation.

The photograph *fig. 1* shows a 4 MeV medical linear accelerator designed and built at the Mullard Research Laboratories, Salfords (England), to Ministry of Health specification. The machine was installed in Newcastle General Hospital during August 1953 and has been in use for the treatment of patients since December 1953. Since the X-rays generated by electrons in this energy range are emitted chiefly in a forward direction, the electron beam itself must be swung round in order to provide for different angles of irradiation of a patient. This facility has been provided with the Newcastle machine by mounting the accelerator on a large double-ended gantry, the X-ray beam being directed towards the axis of rotation where the patient is

[1]) D. W. Fry, The linear electron accelerator, Philips tech. Rev. **14**, 1-12, 1952/53.

[2]) C. F. Bareford and M. G. Kelliher, The 15 million electronvolt linear electron accelerator for Harwell, Philips tech. Rev. **15**, 1-26, 1953/54.

located. *Fig. 2* may serve to illustrate the general lay-out. A more detailed description of this equipment [3]) will appear in a future issue of this Review.

Fig. 3 gives an impression of an even larger medical linear accelerator producing 15 MeV X-rays. This machine has recently been installed in St. Bartholomew's Hospital Medical School, London, by a team of scientists of the Mullard Laboratories headed by T. R. Chippendale under the direction of P. E. Trier [4]). The accelerator is similar to that already described in this Review [2]). Swinging the whole accelerator round for changing the angle of irradiation is not a practical proposition for a machine of this size. In this case, therefore, the accelerating waveguide is mounted in a fixed horizontal position and the electron beam emerging from it is bent through 90° by a powerful electromagnet before hitting the target. The X-ray head housing the target and the electromagnet can be rotated about the horizontal axis of the machine to enable the angle of irradiation to be varied.

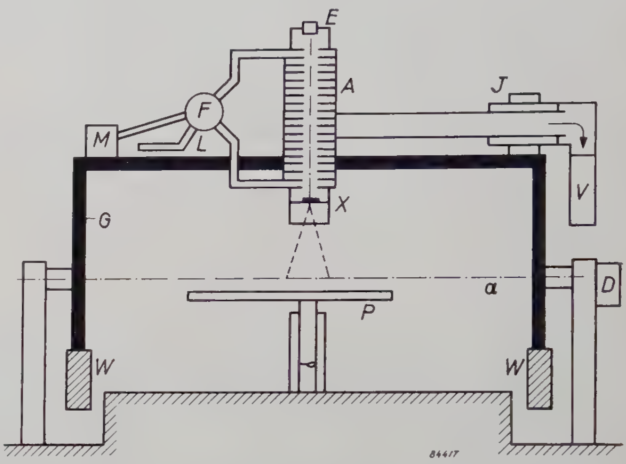

Fig. 2. Schematic representation of the machine shown in fig. 1. *A* is the accelerator proper (corrugated waveguide 1 metre long), mounted in gantry *G* rotatable about the axis *a* by means of an oil motor *D*. *E* electron gun, *X* X-ray head, *M* magnetron, *F* feedback bridge, *V* diffusion pump, *J* rotating vacuum joint, *W* counterweights, *P* patient's couch, *L* water load.

[3]) A preliminary account was given by T. R. Chippendale and M. G. Kelliher, A linear accelerator for X-ray therapy, Discovery **15**, 397-404, 1954 (No. 10).

[4]) A 15 MeV linear accelerator for medical use, Electronic Engineering **26**, 527-528, Dec. 1954 (No. 12).

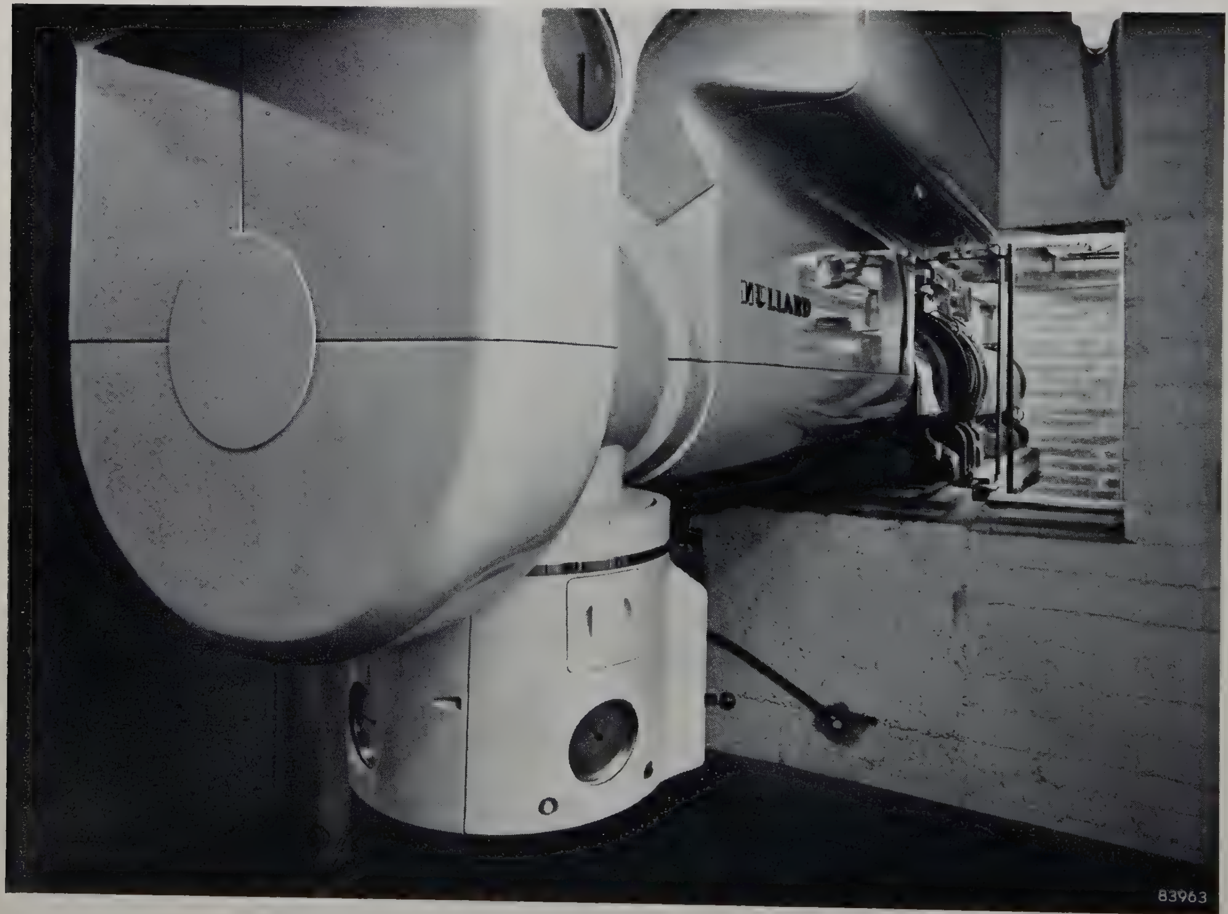

Fig. 3. 15 MeV linear accelerator installed in its temporary building at St. Bartholomew's Hospital Medical School, London.

In fig. 3 it is seen that the target end and X-ray head of the machine are supported from above to permit the positioning of patients underneath. The angular position of the X-ray head is indicated by the scale at top centre. The X-ray beam field size, which may be adjusted using the control buttons on the head, is indicated on the two lower scales.

Four of the six metres length of the accelerator are housed in a separate room which is divided from the treatment room by a concrete wall 1 metre thick. The gap in the wall visible in the phtograph and permitting a complete view of the accelerator is normally closed with concrete blocks.

T. R. CHIPPENDALE.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk * can be obtained free of charge upon application to Philips Electrical Ltd., Century House, Shaftesbury Avenue, London W.C. 2.

2161: R. Vermeulen: Sound recording. General review, presented at the first I.C.A. Congress on Electro-accoustics, Delft, 1953 (Acustica **4**, 17-21, 1954.)

General considerations on sound recording. The following topics are dealt with; improvements in gramophone records, gramophone recording and reproduction, tape recorders; information theory in relation to reproduction of music, fidelity of the reproduction, stereophony, artificial reverberation.

2162: J. Rodrigues de Miranda: The radio set as an instrument for the reproduction of music (Acustica **4**, 38-41, 1954).

In connection with the reproduction of music by radio the following desiderata are put forward: 1. Distortion should be decreased as the high frequency range increases. 2. Means are desirable for adjusting bass response, for cutting off and for gradual attenuation of treble. 3. The electro-acoustical engineer and the architect should cooperate closely. The cabinet must be rigid, the loudspeaker placed forward; its cloth chosen carefully. 4. The speaker should be chosen in accordance with the acoustical properties of the cabinet. The audio-frequency characteristic of the set should be carefully determined.

2163: W. K. Westmijze: Application of the reciprocity theorem to magnetic reproducing heads (Acustica **4**, 50-52, 1954).

It is shown that the output of a magnetic reproducing head can be calculated if the field distribution in front of the head is known for the case when the latter is energized. Application of this method enables us to predict the response curve of a wide-gap head and to explain the existing difference between wide-gap and narrow-gap measurements.

2164: A. G. Th. Becking and A. Rademakers: Noise in condenser microphones (Acustica **4**, 96-98, 1954).

It is pointed out that the mechanical resistance, used for damping a simple mechanical oscillator, acts as the source of a fluctuating force with spectral intensity 4RkT. In the case of a condenser microphone this results in a noise pressure on the diaphragm, which lies in the region of audible sound pressures. Experimental results, obtained with two microphones, agree well with the theory.

2165: A. G. Th. Becking: Methods of measurement for hearing aids (Acustica **4**, 143-145, 1954).

The different ways of defining the input sound pressure on a hearing aid are discussed. In addition, a comparison is given of the specifications for acceptable hearing aids in some European countries.

2166: J. Volger, J. M. Stevels and C. van Amerongen: The dielectric relaxation of glass and the pseudo-capacity of metal-to-glass interfaces, measured at extremely low frequencies. (Philips Res. Rep. **8**, 452-470, 1953, No. 6).

See **R 232**.

2167*: H. Bremmer: On a phase-contrast theory of electron-optical image formation. (Electron Physics, Circular 527, Nat. Bur. Stand., March 1954, 145-158).

The scattering of electrons by an object, as applied in electron microscopy, can be dealt with by means of an integral equation, which follows from the Schrödinger equation corresponding to the electrostatic potential field inside the object. This integral equation can be solved by a series, the consecutive

terms of which may be interpreted as the contribution due to repeated scatterings of the incident electron waves by the object. Summation of the main contribution (for very small electron wavelengths) of these terms leads to a wave function in the object plane (this is a plane immediately behind the object), which in every point depends only on the potential distribution along the straight line joining this point with the electron source. From the values of the wave function in the object plane it is possible, by well-known mathematical methods, to derive the behaviour of the wave function in the space behind that plane. In this way it is possible, in principle, to investigate also the image formation.

2168: J. Haantjes: Die Fernsehübertragung der englischen Kröningsfeier nach dem Festland (Fernmeldetechn. Z. 7, 129-133, 1954, No. 3). (Television transmission of the English coronation festivities to the continent; in German.)

See Philips techn. Rev. **15**, 297-306, 1953.

2169: A. M. Kruithof and A. L. Zijlstra: Les propriétés élastiques d'un certain type de verre (Verres et Réfractaires **8**, 1-13, 1954, No. 1). (The elastic properties of a certain type of glass; in French.)

The disappearance of internal stresses in glass at constant temperature, turns out to be dependent on three factors: the viscosity, the instantaneous elasticity and the elastic after-effect. The first part of the article deals with these factors for a certain type of glass. Special attention is paid to the elastic after-effect. This can be described by a so-called delayed elongation which, after the sudden application of a constant stress, can be represented as a function of time by the sum of two exponential factors.

In the second part, a formula is derived for the disappearance of internal stresses in a rod of stabilized glass at constant temperature, which is suddenly stretched to a specified exent by the application of a constant tensile force. As a result of the mutual effects of viscosity, elasticity and elastic after-effect the resultant stresses disappear. Their disappearance can be described by the sum of three exponential functions. This formula is tested experimentally. The results agree very well with the theory.

2170: E. J. W. Verwey: Das Kräftespiel zwischen Teilchen in lyophoben Kolloidsystemen (Kolloid Z. **136**, 46-52, 1954). (The effect of forces between particles in lyophobic colloids; in German.)

It appears from the theory that the behavior and particularly the stability of lyophobic colloids is in accordance with the balance between the Van der Waals-London forces of attraction and the repulsion forces of the electrolytic double layer. The nature of these forces is examined more closely. The results of the theory point to an elegant explanation of the well known Schulze-Hardy rule, and thereby a further extension of our knowledge of van der Waals-London forces.

2171: H. J. G. Meyer: Theory of radiationless transitions of F centres (Physica **20**, 181-182, 1954).

Calculations are reported, based on the Huang-Rhys model, on the probality of radiationless transitions in F centres. By using a very accurate approximation formula for modified Bessel functions of large index and large argument, an expression is derived which allows a simple physical interpretation for the cases of high and low temperature and which is easily calculated numerically. It is shown that in principle with this model large probabilities for radiationless transitions can be obtained.

2172: J. Volger: Electrical properties of ceramics, Part I. Semiconductors (Research **7**, 196-203 1954).

A review is given of some ceramic products of value in the electrical industry, most of the materials considered being oxides of the transition metals. Some introductory remarks on their crystal structures and on the occurrence of certain substitutions and imperfections in the lattice are followed by a discussion on their electronic conductivity and in particular on the relation between the conduction mechanism and the occurrence of certain transition metal ions in different valency states. The influence of the polycrystalline nature of these materials on conduction and on a number of second-order conduction effects (e.g. Hall effect and frequency dependence of resistivity and magneto-resistance) is briefly discussed. Some applications of ceramic oxidic semiconductors are mentioned.

2173*: C. J. Bouwkamp: Diffraction theory (Rep. Progr. Phys. **17**, 35-100, 1954).

A critical review is presented of recent progress in classical diffraction theory. Both scalar and electromagnetic problems are discussed. The report may serve as an introduction to general diffraction

theory although the main emphasis is on diffraction by plane obstacles. Various modifications of the Kirchhoff and Kottler theories are presented. Diffraction by obstacles small compared with the wavelength is discussed in some detail. Other topics included are: variational formulation of diffraction problems, the Wiener-Hopf technique of solving integral equations of diffraction theory; the rigorous formulation of Babinet's principle, the nature of field singularities at sharp edges, the application of Mathieu functions and spheroidal wave functions to diffraction theory. Reference is made to more than 500 papers published since 1940.

2174: J. Volger: Electrical properties of ceramics, Part II. Dielectries and ferromagnetics (Research **7**, 230-235, 1954).

After some remarks on ceramics as insulating materials, the use of ceramic dielectrics for use in condensors is discussed. The factors favouring a high value of the dielectric constant are considered and in particular the occurrence of such high values with $BaTiO_3$. A number of interesting properties of soft ferromagnetic ceramics than are given, all of the ferrite or ferroxcube type. The saturation magnetization values are discussed in connection with the theory of *Néel*. Attention is given to the h.f. properties of ferroxcube, i.e. the various resonance and relaxation phenomena which occur. Finally, a hard ferromagnetic ceramic called magnadur (also known as ferroxdure) is dealt with. This is a hexagonal oxidic compound of iron and barium with an extremely large coercive force. It may be used for permanent magnets which must withstand large demagnetizing fields.

2175: R. Vermeulen: Stereophonic reproduction (Audio Engineering **38**, 21, 1954, No. 5).

Recapitulation of the work of K. de Boer on the mechanisms of binaural and stereophonic sound phenomena. The perception of direction is discussed as being due to differences both in the time of arrival of the sound and in the intensity. The differences between binaural and stereophonic listening are outlined; in the latter case the sound image is less well defined. It is pointed out that the aim is not to try to reconstruct the original sound field, but to deliver the correct sound to each ear to simulate a sound source from a certain direction. Some empirical rules for the placing of microphones and loudspeakers are given. Finally mention is made of the part played by head movements in the location of the sound image by the listener.

2176: J. L. H. Jonker: Ten-volt effect with oxide-coated cathode (Electronic Engineering **26**, 282, 1954).

According to a theory given earlier, the anomaly in the diode characteristic that occurs at an anode potential of 10 V should be attributed to the fact that the space-charge contribution of electrons reflected from the anode then shows a minimum (see Philips Res. Rep. **2**, 331-339, 1947). This theory is confirmed by measurements on the reflection and secondary emission of a surface that had been maintained for some time facing an emitting oxide cathode.

2177: B. Combée and A. Engström: A new device for micro-radiography and a simplified technique for the determination of the mass of cytological structures (Biochim. et Bioph. Acta **14**, 432-434, 1954).

A sealed-off X-ray tube with a very thin Be window is used as a source of soft X-rays for quantitative historadiology by means of contact micro-radiography. A procedure is described whereby the mass of the specimens may be determined without the necessity of reference exposures. The resolution of the image obtained with this tube is about 0.5 μ.

2178: H. de Lange: Relationship between critical flicker-frequency and a set of low-frequency characteristics of the eye (J. Opt. Soc. America **44**, 380-389, 1954)

Measurements of the critical flicker frequency of the eye as a function of both the average luminance of the test field and the time variation of this luminance were recorded by plotting the "ripple ratio" r against the critical frequency; r is defined as (amplitude of first Fourier component)/(average luminance) of the stimulus. It is shown that with constant average luminance, the points observed for various time functions fit into one smooth curve, which for low luminance is monotonic. At high luminances the curve shows a minimum in r at a critical frequency of about 9 c/s. This means that the eye has a maximum sensitivity to flicker at this frequency.

2179: L. A. Æ. Sluyterman and B. Labruyère: Side reactions in the polymerization of α-amino acid N-carbonic anhydrides. Titration data of polyglycine and polyalanine (Rec. trav. chim. Pays-Bas **73**, 347-354-1954).

Polyglycine and polyalanine obtained by polymerization of their N-carbonic anhydrides often

appeared to certain more acid groups than animo groups. This is ascribed, at least partly, to the formation of hydantoin groups. A small amount of 2.5-diketopiperazine was isolated from polyglycine. A discussion of the results and their bearing upon the calculation of molecular weights is given.

2180: F. A. Kröger, H. J. Vink and J. van den Boomgaard: Controlled conductivity in CdS single crystals (Z. phys. Chem. **203**, 1-72 1954, No. 1/2).

The electrical properties of single crystals of pure CdS or of CdS containing gallium, indium, antimony, chlorine or silver depend markedly on the atmosphere of preparation. Crystals subjected to an oxidizing atmosphere (i.e. sulphur vapour) are insulators or semi-conductors, showing photoconductivity; crystals subjected to a reducing atmosphere show quasi-metallic, electronic conductivity, For crystals doped with Ga or C1, the number of carriers in the reduced crystals is constant over a wide range of temperatures and equal to the concentrations of foreign ions over a wide range of atmospheres (controlled valency). The optical properties (absorption, fluorescence) vary also with the atmosphere, absorption bands in the yellow part of the spectrum appearing in oxidized, but not in reduced crystals. A general theory is developed by extending Schottky and Wagners theory of lattice imperfections along the lines indicated by Schottky (1935), By means of this theory it is possible to calculate the dependence of the concentration of the various kinds of lattice imperfections, donors, traps, and acceptors, on the concentration and nature of the impurities and the reducing power of the atmosphere. Applying this theory to CdS, a satisfactory agreement with the experiments is obtained.

2181: J. te Winkel: Lijnversterkers voor draaggolfsystemen op coaxiale kabels (De Ingenieur **66**, E.61-65, 1954, No. 25). (Line amplifiers for carrier telephone systems on coaxial cable; in Dutch.)

A brief survey of the principles governing the design of these amplifiers.

2182: E. Havinga and J. P. L. Bots: Studies on Vitamin D, I. The synthesis of vitamin D_3 3 C^{14} (Rec. trav. chim. Pays-Bas **73**, 393-400, 1954).

In connection with the study of the metabolism of vitamin D, vitamin D_3 has been synthesised with a C^{14} atom in place of the third carbon atom. The preparation of this vitamin D_3 3 C^{14} is described. Special attention is paid to the technique of the photochemical conversion of 7-dehydrocholesterol 3 C^{14} (small quantities) and to the isolation of the labelled vitamin D_3 from the irradiated mixture. Some biological experiments are briefly described.

2183: G. W. Rathenau and G. Baas: Croissance des grains, observée par microscopie electronique à emission (Métaux, Corros. Ind. **29**, 139-150, 1954 (No. 344). (Grain growth observed by means of the emission electron microscope: in French.)

Changes in structure in metals and alloys at elevated temperatures can directly be observed applying emission electron microscopy. With the instrument used for these investigations a resolving power of about 1000 Å is obtained.

Rolled face-centred cubic NiFe alloys have been used to study grain growth in a texture. This investigation includes growth of one large grain of deviating orientation in a well-pronounced texture as well as grain growth in an imperfect texture. In both cases only high energy boundaries move at measurable speed. The results are discussed in terms of interfacial tension effects.

An example of grain growth hindered by inclusions is given for a Cr Ni steel specimen, containing inclusions of σ-phase.

Grain growth accompanying the transformation $\alpha \rightarrow \gamma$ in a SiFe alloy has been described. In this case the rate of diffusion determines the rate of growth of the γ grains.

The austenite-pearlite transformation in eutectoid carbon steel has been directly observed. The existence of an orientation relationship between austenite and pearlite is improbable as shown by the observations. Grain-boundary displacements in the austenite have been observed near a growing colony of pearlite.